ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies

SUBMARINE CABLES IN NORWAY
MINE DETECTORS AND MINE DISPOSAL IN FRANCE
EQUIPMENT FOR NEW OSLO BROADCASTING HOUSE
PORTABLE CRYSTAL-CONTROLLED FREQUENCY STANDARD
REMOTE CONTROL
CONTROL STATION APPARATUS
VERY-HIGH-FREQUENCY TRIODE CIRCUITS
LINEAR THEORY OF BRIDGE AND RING MODULATORS
ACCURACY OF IMPEDANCE MEASUREMENTS
HEATING OF RADIO-FREQUENCY CABLES

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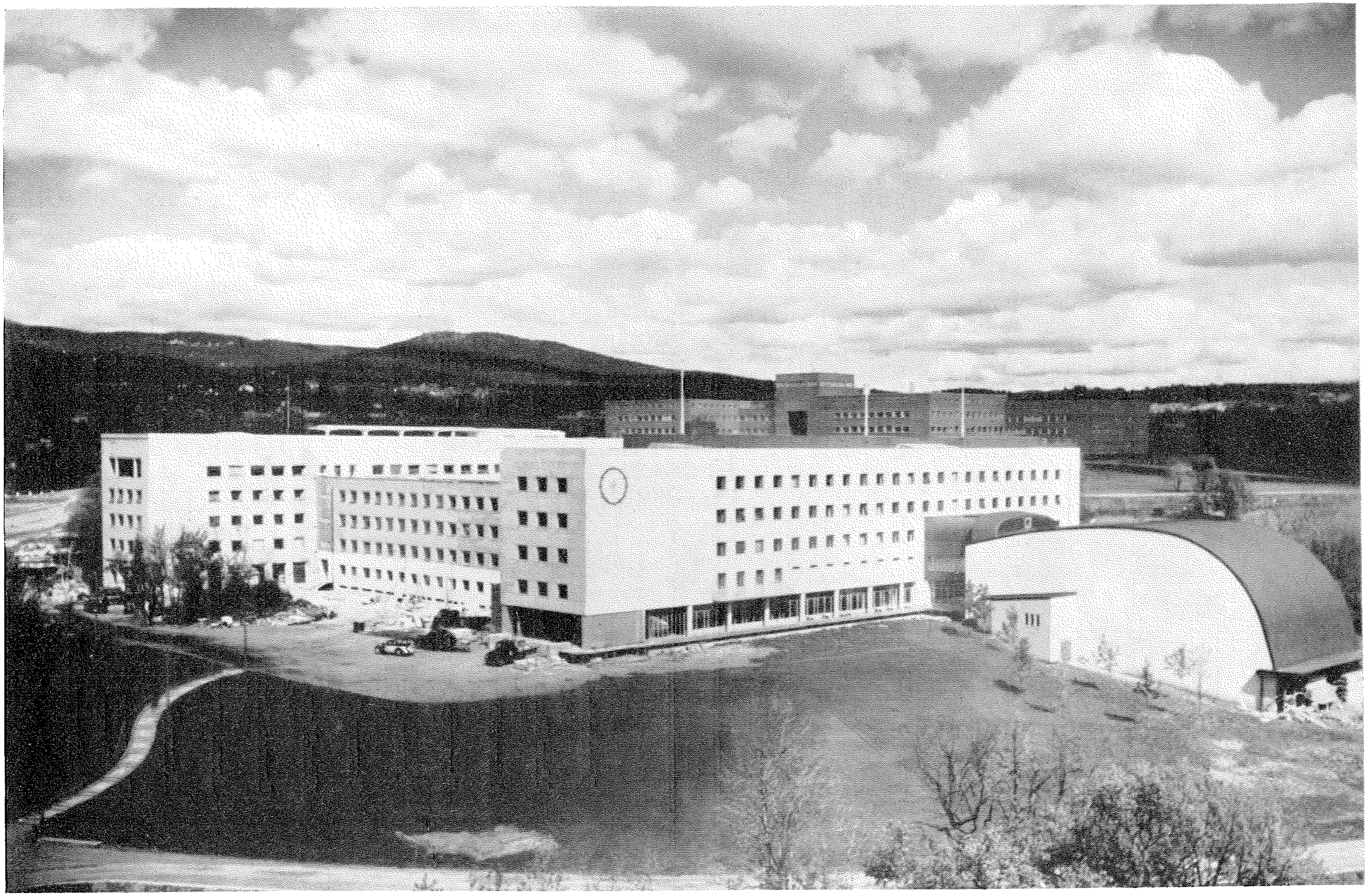


Broadcasting House at Oslo, Norway. The great concert studio with its curved roof is at the extreme right. The main group of studios is housed in the wing of the building adjacent to the concert studio.

Submarine Cables in Norway

By CHRISTIAN W. HIRSCH

Standard Telefon og Kabelfabrik A/S, Oslo, Norway

THE Norwegian coastline, measured around all islands and fiords, has a total length of about 20,000 kilometers (12,427 miles). Under the influence of the Gulf Stream, the coastal climate is rather mild, in spite of the northerly position of the country, between 58 and 71 degrees north. Fisheries and shipping are important sources of income, and a comparatively great part of the population lives along the coast, even in the roughest and most isolated parts.

Electrification Progress

The accompanying map shows the fiords which penetrate far into the mainland, while the inner coastline is protected against the open sea by numerous islands—the Norwegian skerries. The fiords are deep, and there are a great number of waterfalls, which can be harnessed cheaply for the production of electric energy. These conditions, in conjunction with convenient ice-free harbors, are very suitable for the large-scale development of an electrochemical industry.

The great industrial plants along the coast usually supply the neighboring districts with electric power and, in addition, many communities, counties, and cities have built their own stations. However, there are still numerous districts which as yet have no supply of electric power. The distances over which energy must be transmitted may be fairly long, since the population is quite scattered, and the distribution costs are, consequently, rather high because of the comparatively low demand. Therefore, over wide areas along the coast, and particularly in the north, the population has not yet been able to raise the funds necessary to secure a supply of electric power.

During the years immediately preceding and those following the war, the Norwegian government subsidized the electrification schemes in these districts; partly to promote industry and fish-refrigerating plants and partly to provide electric power for cooking and heating, thus avoiding excessive cutting of timber in the coastal districts, or the excavation of peat for fuel in areas where the ground could otherwise be quite easily cultivated

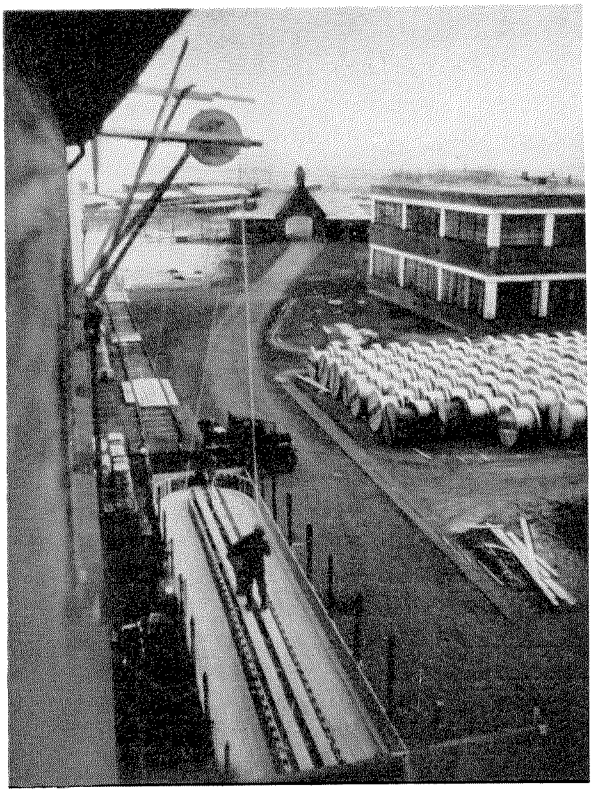
The electrification of the coastal districts necessitates a comprehensive use of submarine cables between islands and from islands to the mainland, and also for crossing the fiords which cut into the mainland. Power transmission is normally at 20 kilovolts, with 3-phase, 50-cycle-per-second supply. This is also generally used for transmission over submarine cables. These cables are of the 3-core, 25- or 35-square-millimeter-conductor *H* type, being lead covered with single-steel-wire armoring and jute serving. The armoring is, of course, specially designed to withstand the very high stresses occurring during laying the cable and in subsequent service.

As the 20-kilovolt cable can carry 4000 kilovolt-amperes, and as the load in many cases is only 10 percent or less of this rated capacity, the voltage for secondary supply mains connected to the main cable line is sometimes reduced to 10 kilovolts, so that ordinary belt-insulated submarine cables may be employed. Only a very simple calculation is necessary to determine if this is economically feasible. Another alternative is the use of aluminum conductors instead of copper.

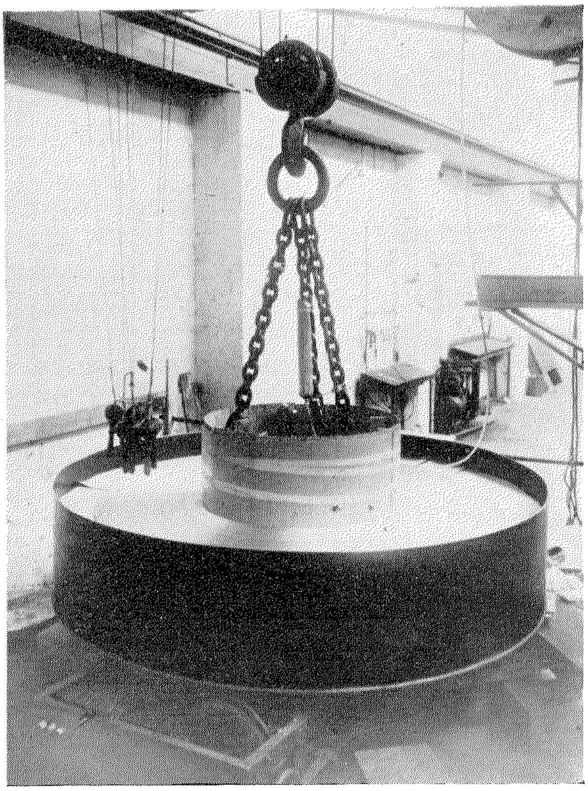
During and after the war, several 60-kilovolt submarine cables have been designed and ordered for the main supply lines to cities and some of the larger islands. Usually, oil-filled cables are used, to reduce insulation thickness and thereby weight and price. Gas-pressurized cables can also be used. At depths greater than 150 meters (492 feet), the water pressure will exceed the internal 15 atmospheres of gas pressure and the reinforcing tape normally placed around the lead sheath can be omitted.

Sea Water as Return Conductor

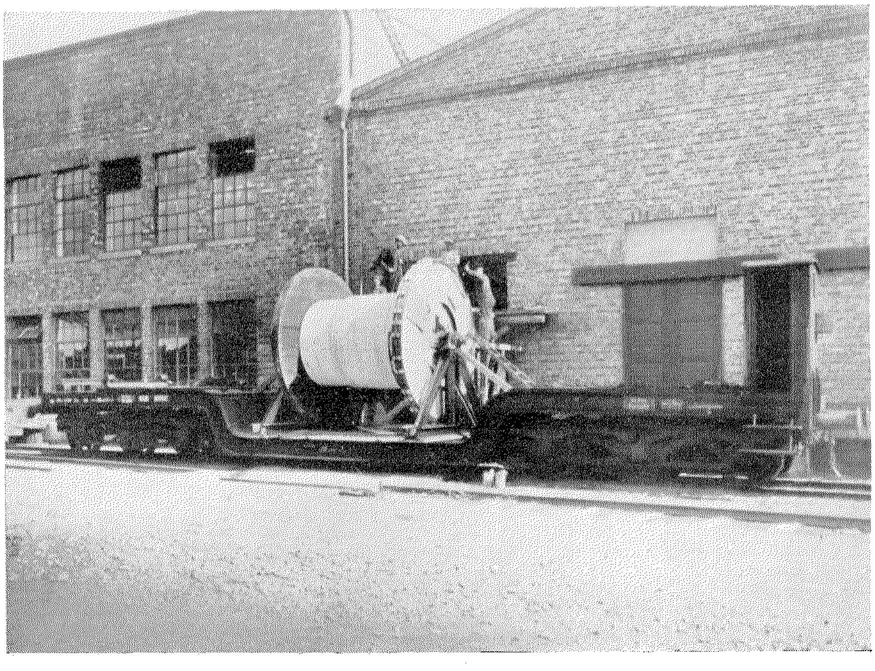
The possibility of single-phase submarine cables using the water for the return conductor has been discussed. Tests on a submarine cable



Cable being coiled on a railway flat car for transportation to the cable ship. Several such cars are necessary when a cable of considerable length is shipped.



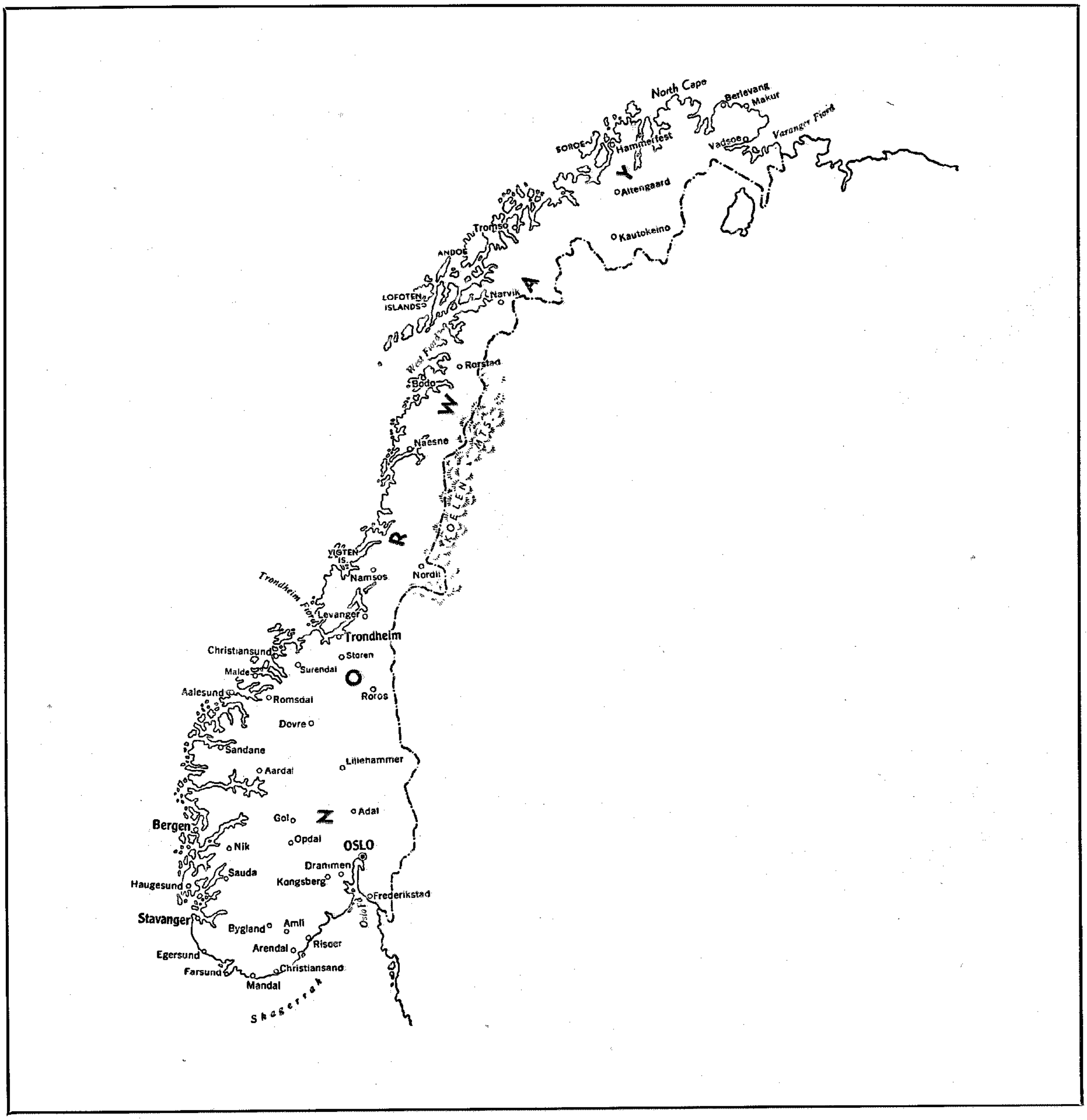
Coiling submarine telephone cable after lead-covering process. The cable has yet to be armored and jute served, and is transported on the above reel for these processes.



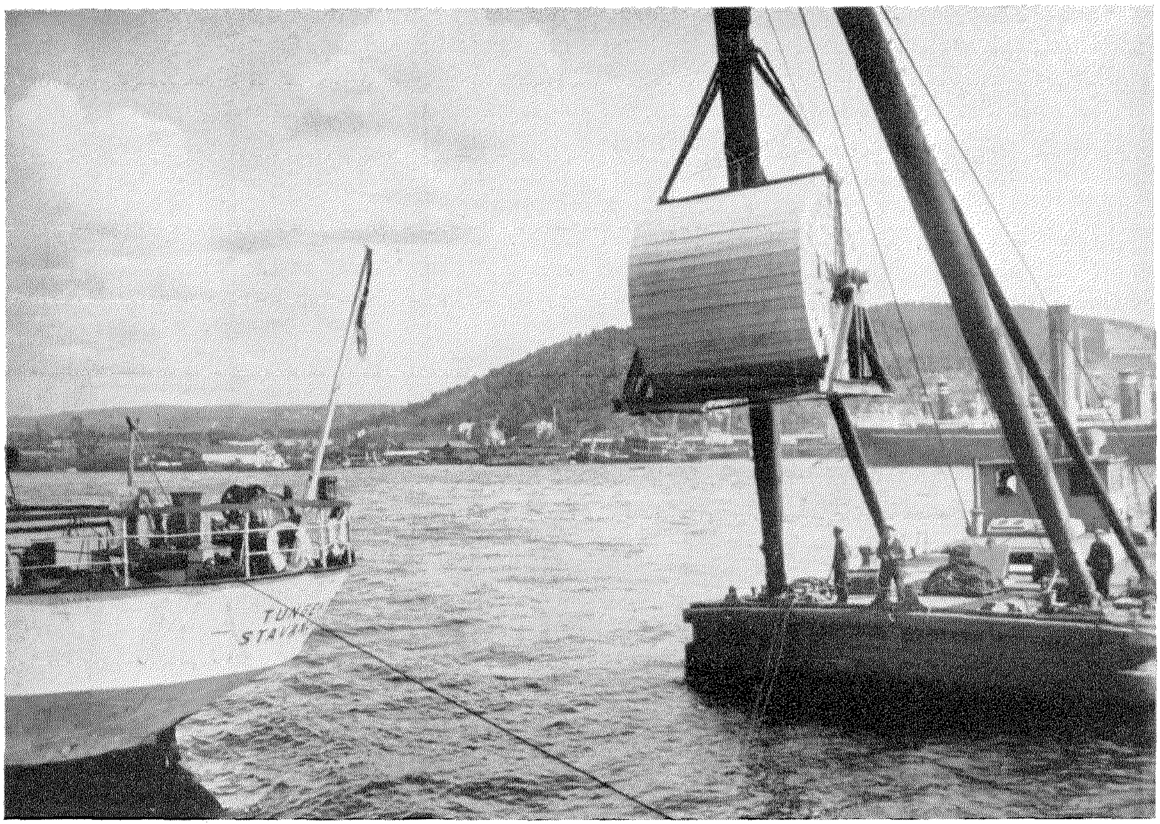
Cable being reeled for shipment on special railway car designed for heavy loading.

showed that salt water, the lead sheath, and the armoring, all in parallel, have a resistance of 0.084 ohm per kilometer (0.156 ohm per nautical mile) at 50 cycles, and the short-circuit impedance has a phase angle of $\cos \varphi = 0.93$.

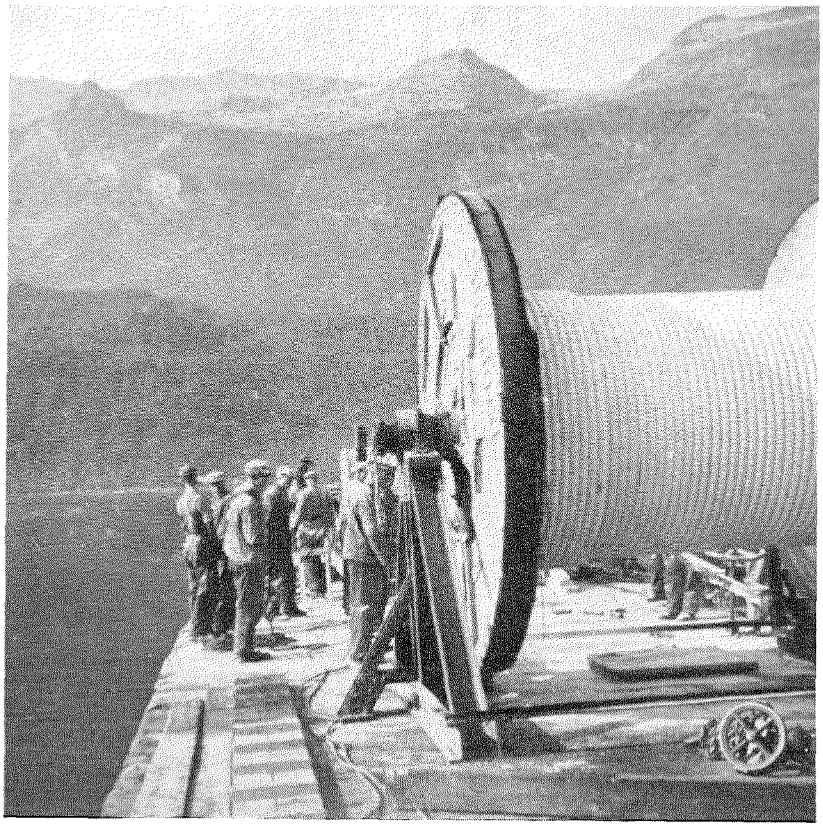
The return conductivity thus represents a copper conductor of 150-square-millimeter cross section, and the voltage drop in a single-phase cable with a conductor of small cross section is not much higher than in a 3-phase cable of the same working voltage and conductor cross section. As the cost of the single-phase cable is less than 50 percent of the 3-phase cable, it is possible to use single-phase cables for many projects which otherwise would be too expensive. A special transformer is required, but this does not add to the cost, as the voltage would have to be reduced anyway in most cases. The lack



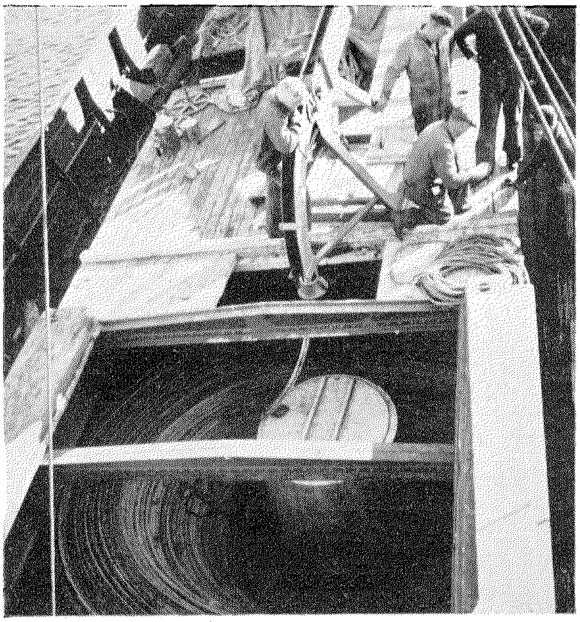
Map of Norway, illustrating the numerous fiords and islands along the coast. Coastwise transmission of power and of telephone service is facilitated by the use of submarine cables.



Loading a 38-ton reel of submarine cable onto a cable ship.



Reel of cable on a cable-laying lighter. The terrain at a typical deep fiord is shown here.



Cable coiled in ship's hold. Note the trough through which the cable is fed from the hold to a trumpet at the bow of the ship, and thence into the sea.

of 3-phase current might be a drawback to the development of industry, and if 3-phase current should eventually be required, another single-phase cable may be laid and the water used for the third conductor.

Important main-line cables can be constructed with three single-phase cables, and if one of the cables should break down, the water can be used as a spare conductor until repairs are made.

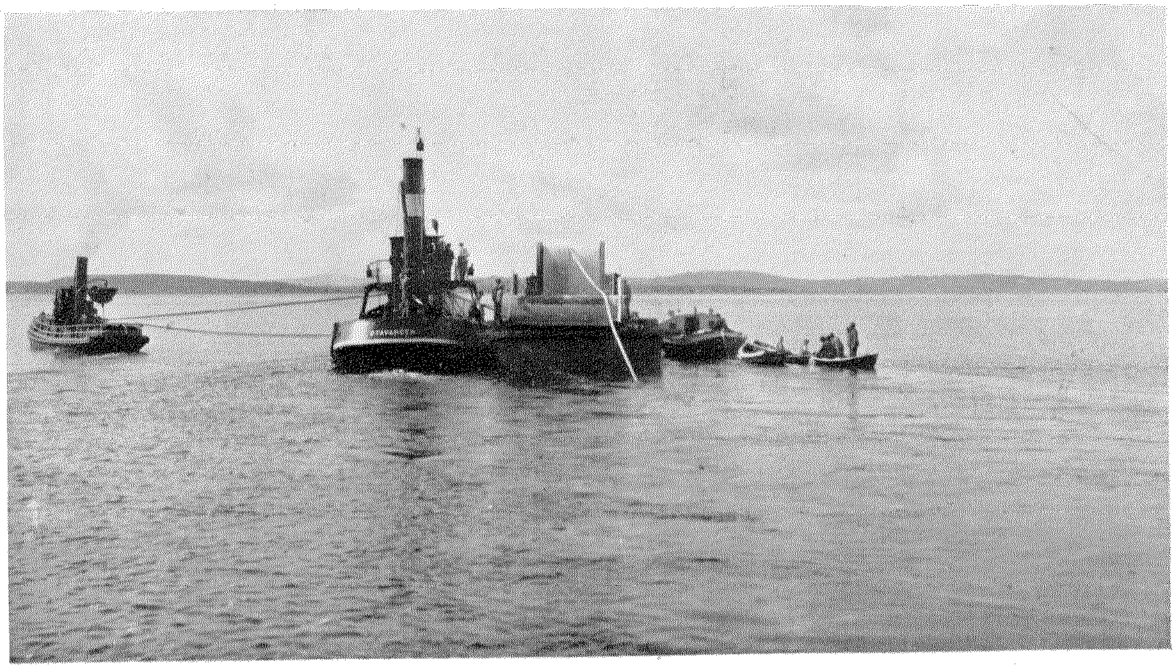
The impedance of salt water increases very rapidly with frequency. At 800 cycles nearly the whole return current will flow in the lead sheath. This is important, because it proves that disturbances in a single-phase submarine power cable will not affect neighboring telephone cables.

Manufacture

Standard Telefon og Kabelfabrik A/S, the only Norwegian manufacturer of lead-covered cables, has manufactured and installed most of the submarine power cables in Norway for many years. During 1939, one year before the German attack, more than 150 tons of submarine power cables were manufactured, and 9 cables were laid on the north and west coasts.

The manufacture, transportation, and installation of submarine cables raises many interesting technical problems. Within the factory, the problem is covering very long lengths of cables with lead, and the necessary transportation

Laying of submarine cable from a lighter. Tug boats are used to move the lighter.



between fabrication processes (see accompanying photographs). Submarine power cables in single lengths up to 6 kilometers (3.24 nautical miles) are not uncommon, and the weight of a finished length of such cable amounts in many cases to more than 50 tons. Submarine telephone cables are often even longer and heavier. In May, 1947, manufacture of a deep-sea telephone cable in one length of 20 kilometers (10.8 nautical miles) was completed. The total weight of this cable was 125 tons, and it occupied 3 adjacent railway cars for transportation from the factory to the harbor. The cable was constructed with 3 paper-insulated star quads of 1.4-millimeter copper conductors, each quad shielded. The capacitance is approximately 0.04 microfarad per kilometer (0.073 microfarad per nautical mile). The twisted quads were protected against outside water pressure by a spiral steel tape within a 3.0-millimeter lead sheath (2-percent tin alloy). The armoring consisted of 4.2-millimeter galvanized-steel wires with outside jute serving and a total diameter of 40 millimeters.

This cable was manufactured for the Norwegian telegraph administration, and is to be laid in depths as great as 800 meters (2625 feet).

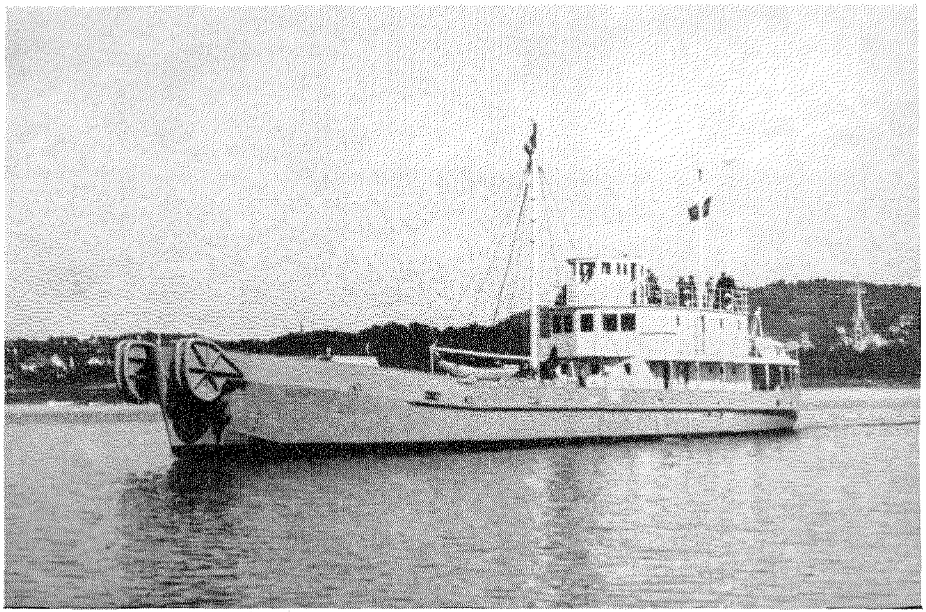
Transportation

Hitherto, the main difficulties in the transportation of submarine cables have been the long distances and the lack of lifting cranes of sufficient capacity in the remote coastal districts. Laying the cable can be done either from a reel on a lighter, or by means of specially equipped cable ships. The method chosen depends on the conditions.

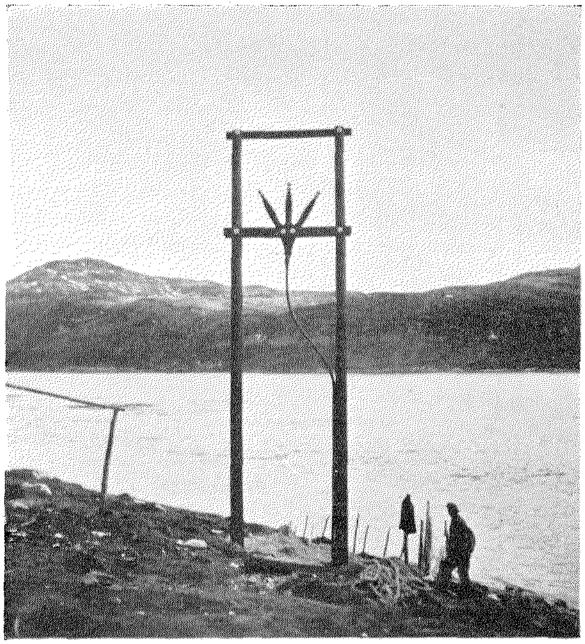
If the cable should be laid not too far from Oslo, or wherever lifting facilities were adequate, the cable could be coiled on a big reel, installed in a lighter specially fitted with a reel-brake equipment, and towed to the site. The reel was transported by railway to the most convenient harbor. If the cable was very heavy or long, or the towing distance too great, a special cable ship was used. To reduce costs, several lengths of cable were usually transported and laid during each trip. In this case, the cables were coiled in the ship's hold and paid out over a bow wheel by means of a capstan device, an arrangement no doubt familiar to many. The photographs included with this article illustrate some of these methods.

In the future, most of the submarine cables manufactured by Standard Telefon og Kabelfabrik are expected to be transported and installed by means of the company's new cable ship *Stanelco*, which was put in service last summer.

Stanelco is a former antiaircraft gunboat, originally built in England in 1944 for landing operations and reconstructed in Norwegian shipyards. The ship is 154 feet long, 21 feet wide,



The cable ship *Stanelco*.



Cable installation at 71 degrees north latitude. The pot-head termination of the 3-phase cable is supported on the lower wooden cross arm.

with 2 cable store rooms of 5600 cubic feet. It has a displacement of 310 tons. The machine room is equipped with two diesel engines, each of 500 horsepower, and three diesel-driven direct-current electric generators, each rated at 30 kilowatts, 220 volts. *Stanelco* has many facilities that make the ship well fitted for cable laying, such as sounding equipment and a portable public-address system. Furthermore, a 40-watt duplex radiotelephone installation makes it possible to communicate with the office in Oslo regardless of the ship's position.

Installation

The laying of submarine cables always involves a certain amount of uncertainty, and requires experience and careful planning. The Norwegian fiords are bounded by high and steep mountains which extend in many cases almost vertically below the surface of the water to the bed of the fiord, while the bottom itself is more or less flat. The depths vary in most cases from 200 to 300 meters (656–984 feet), but many fiords are more than 600 meters (1969 feet) deep, and some even exceed 1200 meters (3937 feet). As the cable generally weighs 10 to 15 kilograms per meter (20 to 30 pounds per yard), it will be subject to very high stresses during the paying out, and also at the shore ends where the cable is anchored. It is most important during the laying process to give the cable at the bottom sufficiently high tension to avoid coiling, but it must not be so high that the cable does not closely follow the bottom profile. At the greater depths this is rather difficult to do.

Out in the skerries, the sea is generally shallower than in the fiords. Unfortunately, however, the placing of submarine cables is impeded here by strong tides and heavy seas from the ocean. When in service, the cables in the skerries are greatly exposed to storm and damage from breakers, and must be carefully protected where they are brought ashore; furthermore, specially heavy armoring must be applied where the water is shallow, to make the cable secure from damage resulting from severe ground swells.

The submarine cables are normally installed during the summer season, but it may happen that the work must be done in the autumn, and even in the wintertime. In such cases, one must allow for long delays due to storm and rough weather. In October, 1939, during an installation near Hammerfest, the most northern city in the world, the cable ship was weatherbound for one week in a snowstorm, and at another installation south of Bodö in December, 1939, the transportation of the cable reel by sea was delayed for one week by a snowstorm. Finally the cable arrived and was laid on the 20th of December in calm, bright weather; at this time of the year there are only 4 hours of daylight in this area, which is within the arctic circle. The next day a storm blew up which continued for three weeks.

The low temperature gives trouble in the wintertime: In January, some years ago, a high-tension submarine cable was transported by railway to Narvik. This was during an extremely cold period with the temperature at approximately −60 degrees Fahrenheit. Although this does no harm to the cable, it becomes as stiff as wood and cannot be handled. When received in Narvik, the cable was heated electrically, a process which continued as it was paid out to the cable ship.

However, as mentioned above, submarine cables are normally installed in the summer, when it is an interesting and agreeable job to people who like salt water.

Mine Detectors and Mine Disposal in France

By ANDRÉ VIOLET

Le Matériel Téléphonique, Boulogne sur Seine, France

DETECTION of land mines was of great importance during World War II and after the termination of hostilities. Several types of mines that were used in large numbers are described briefly. The operating principles of the most successful mine detectors are outlined, and descriptions are given of instruments produced in Great Britain, France, Germany, and the U.S.A.

• • •

Mines were first used extensively in the 1914–1918 war. In addition to sea mines and land mines under trench supports, shells, torpedoes, and grenades fitted with contact fuses were carefully concealed in the ground. The purpose of mines is usually to slow down an advance, with maximum losses for the attackers and minimum losses for the defenders. In addition, a psychological demoralizing effect is produced.

During the second world war, mines were employed by all belligerents on a very large scale. In the first weeks of 1939, French attacks were met with various destructive devices; explosions were set off by the sudden movement of a wire stretched close to the ground or by the opening of the door of an empty house. "Booby traps" were built around such things as fountain pens, bicycles, dead horses, and food, used as "bait."

Since that time, the technique of booby traps has advanced in cunning. For instance, an abandoned house was left with its door ajar to invite investigation; the mine remover, taking cover behind a barrel in front of the house, pushed the door open with a pole and was struck from behind by a mine set up in a ditch and connected to the door by an underground wire.

1. *German Mines*

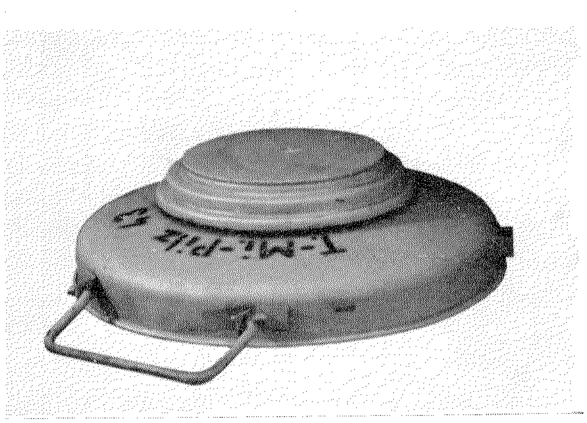
All German mines used in France fall into two classes:

A. Antitank and antivehicle mines, containing several pounds of high explosive, operated by loads of the order of 400 to 600 pounds, and causing damage to or destruction of vehicles, whether armored or not.

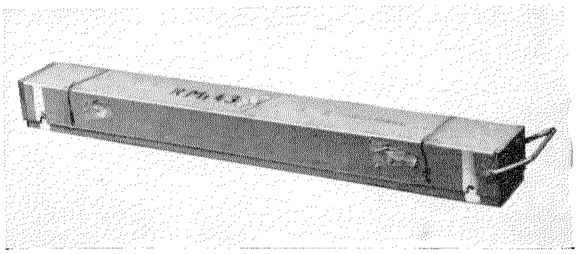
B. Antipersonnel mines, operated by loads varying from a few ounces to several pounds, often causing fatal wounds either by their blasting effects in shattering the lower limbs or by throwing splinters causing deep wounds. Frequently they were in the form of booby traps.

Mines of the first kind were sometimes used with light-pressure fuses; they then fall into the second class. They may even incorporate features in both categories by having a main fuse that is not very sensitive and a highly sensitive secondary fuse.

Among the devices of the first category, that most usually encountered was the *Tellermine*, a circular steel container, 10½ inches in diameter and about 4 inches in height. It contains about 11 pounds of explosive; its fuse is sometimes designed to operate either by being stepped on or by moving the mine. It may also cover a second



***Tellermine*, a steel container holding about 11 pounds of explosive. Normally fused to explode when stepped on. Used for antitank and antivehicle work.**



***Riegelmine* for antitank and antivehicle service. The steel box is about 31 inches long.**

mine, which explodes when unearthed. Illustrations of this mine and others described in this section are included in the paper.

Among other antitank and antivehicle mines, mention must be made of the *Riegelmine*, a rectangular steel box 31 inches long; the *Holzmine*, a wooden mine one foot in length which contains but little metal, except in the form of the fuse, nails, and iron wire; and the *Topfmine*, a large asphalt container with a chemical igniter consisting of sulphuric acid and potassium chlorate, and with practically no metal.

Among the usual antipersonnel mines is the *S* mine, which is 4 inches in diameter and 4¾ inches in height, and contains a bomb loaded with steel balls or metal fragments. When operated by a vertical pressure fuse, normally set for 15 to 25 pounds, the bomb jumps up 6 feet before exploding, and the scattering metal splinters are generally fatal over a radius of about 150 feet. This device was most respected by the mine-clearing squads.

The *Schümine*, is a small wooden foot-shattering mine fitted with a bakelite-and-metal fuse and loaded with 200 grams of explosive.

The glass mine is generally provided with a steel plate and a metal lever-type fuse.

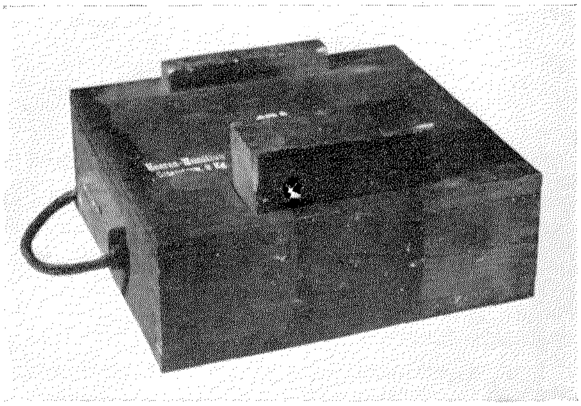
The *mustard pot* is a small sheet-metal mine only 3 inches in diameter.

Additional types of mines made by the various belligerents were also encountered; they include about 150 different types.

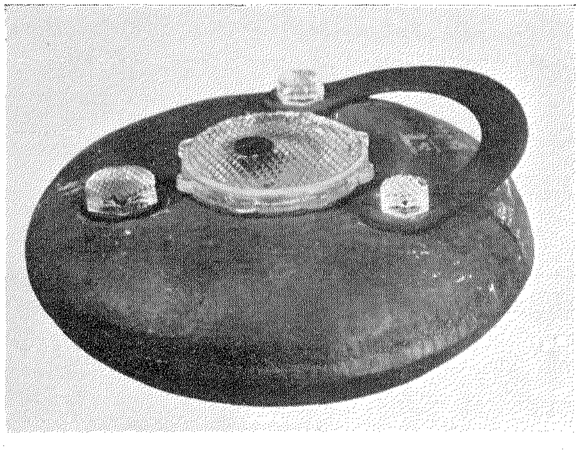
After the liberation, there were in French territory about 15 million mines, generally scattered along the coastal regions and in the Normandy and Alsace-Lorraine battlefields. A certain number were also found around such places as air fields, radio transmitters, service headquarters, and high-voltage towers to give protection against sabotage during the occupation by the enemy.

Those that had been laid by the Germans before active military operations began were arranged in parallel lines or in diamond formation, and were indicated on the German maps handed over after capitulation. Apart from a few errors, apparently accidental, and a few "dummy mine fields," which were areas reported as mined although actually clear, these maps have been of great service for nonmilitary mine-disposal purposes. Unfortunately, information was completely lacking on the "protection" mines laid by the Germans during the battles in Normandy and Alsace, and on booby-traps in the houses in the coastal areas, in Lorraine and Alsace, and in certain mine fields set up for local operations bordering on the Atlantic.

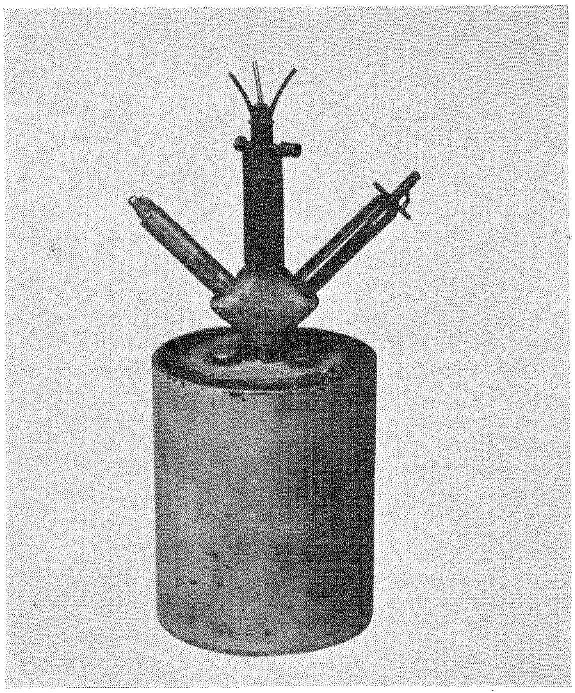
In these cases, it was necessary to use devices capable of destroying or detecting the mines. The first solution is the most satisfactory; it is quicker, and it eliminates accidents during removal and depriming, which even efficient



***Holzmine*, contained in a wooden box about a foot long, was used against tanks and vehicles.**



***Topfmine* contains practically no metal. The case is of asphalt and the igniter is of a chemical nature. This was for antitank and antivehicle use.**



S mine for antipersonnel service. Only 4 inches in diameter, the bomb, when activated by a pressure of about 15 pounds, jumps up 6 feet and scatters metal balls or fragments lethally over a radius of about 150 feet. Some models have but one centrally located fuse.

detection cannot obviate. Particularly effective is the use of an arrangement of chains that strike the ground vigorously in front of a tank and cause mines to explode. Unfortunately, this device is put out of action fairly rapidly. In sandy places, a high-pressure water jet gives better results. A roller can also be used. However, no general solution has yet been found for this problem.

2. *Mine Detection*

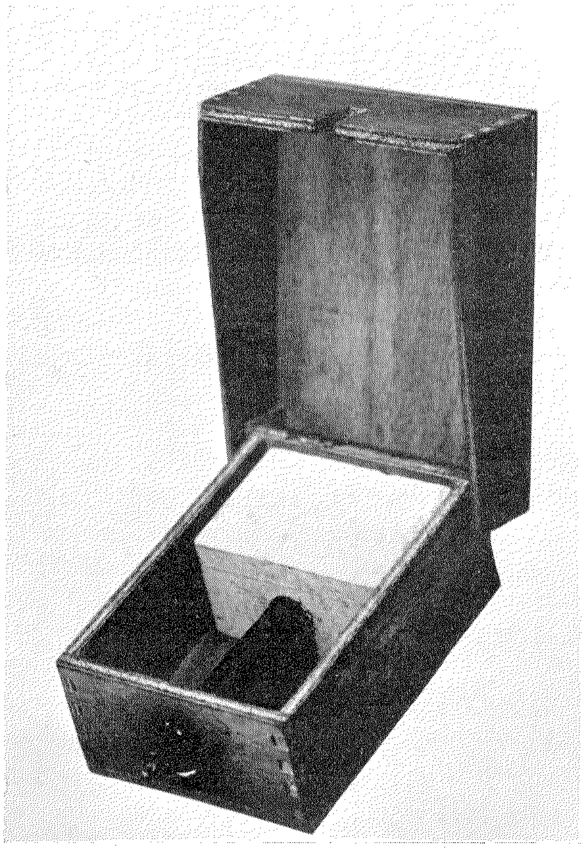
2.1 General Principles

Any buried mine is essentially a foreign body in the ground. The mine can be detected as an obstacle by driving a sufficiently long probe into the earth. This work is slow and tiring, for the ground must be examined inch by inch in suspicious places. It is effective only in loose ground. Accidents have been reported in which the probe strikes the pressure fuse or pin of the mine.

This extremely primitive method is sometimes the only one practicable. It would be more scientific, however, to make use of the irregularities caused by the mine in such characteristics as soil density, electrical resistivity, dielectric constant, acoustical resistivity, susceptibility, magnetic permeability, gravitational pull, the earth's magnetic field, and absorption or reflection of Hertzian waves.

Up to the present, only radio methods, based on reflection or absorption of radiated waves, and electromagnetic methods, operating on variation of self or mutual inductance in nearby electrical circuits, have been utilized.

The first method is absolutely general, but its operation raises so many difficulties that all detectors manufactured in large quantities use the second method, which is applicable to electrical conductors and magnetic objects.



***Schümine*, shown with cover open, was primarily designed for personnel foot shattering. This mine was loaded with 200 grams of explosive.**

2.2 RADIO METHODS

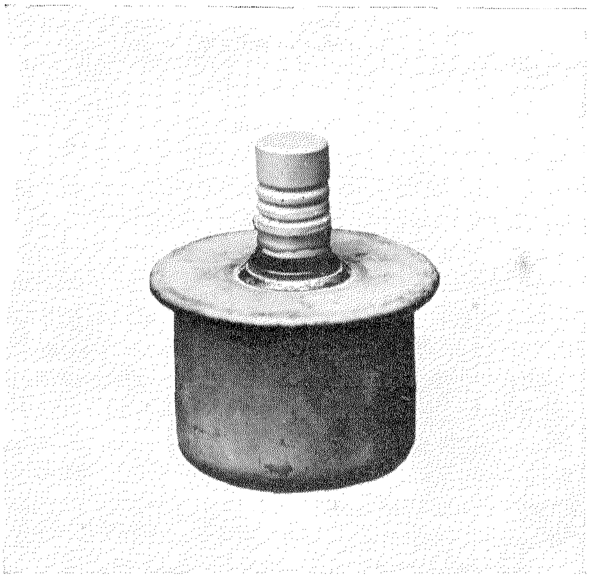
2.2.1 *Principles*

If an antenna that is directly coupled to an oscillator is placed close to the ground, the radiation resistance of the antenna will depend on the dielectric properties of the ground. The behavior of the oscillator will be influenced by these load variations. For example, if the surface of the ground is conducting and reflects the radio waves, the output power increases; if it is absorbing, it decreases. Of course, the distance between the antenna and ground is very important, and for mine detection must be kept constant.

2.2.2 *American Detector AN/PRS*

The circuit of the AN/PRS detector, developed in the U.S.A., is shown in Fig. 1 and is based on the above principle. An acorn tube is used as an oscillator with a line resonating at about 300 megacycles per second (1 meter). It is coupled to a radiating dipole and reflector. The oscillator grid current is measured by a microammeter. The direct component of grid voltage biases a low-frequency amplifier having a 1000-cycle input from a resistance-capacitance oscillator. The transmitter grid current decreases and the note in the headphones becomes more intense as the absorption of the transmitted wave increases.

These waves penetrate but an inch or two into the ground, and the instrument is sensitive only to bodies at least 6 inches in length. Ploughed earth, layers of leaves, dead twigs, pebbles, grass, and humidity affect it. Consequently, the handling of this apparatus is very delicate and only an experienced operator would appear to be able to use it successfully.



***Mustard Pot* mine for antipersonnel use.**

Mechanically, the transmitter, antenna, and reflector are at the end of an adjustable handle on which the microammeter is mounted. A box containing the power supply and audio-frequency equipment is carried on the back of the operator. The whole assembly is somewhat fragile and cumbersome.

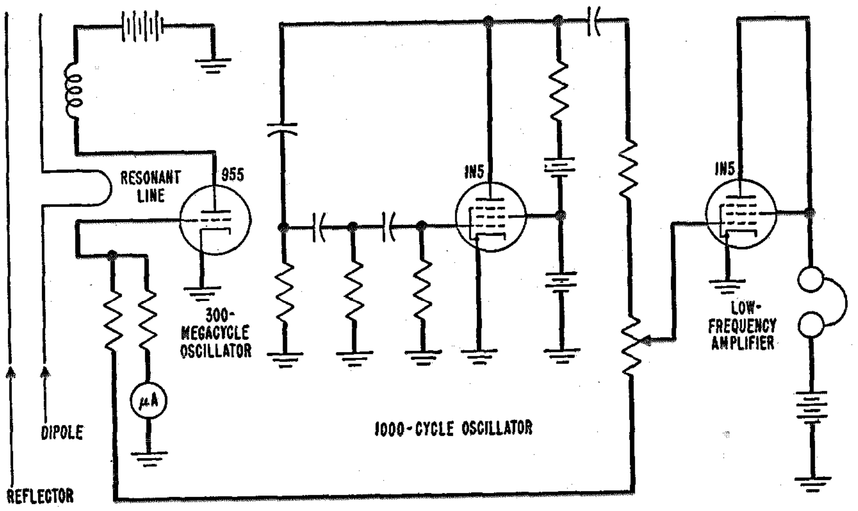


Fig. 1—Schematic arrangement of the AN/PRS mine detector. The 955 oscillates at about 300 megacycles, its performance being influenced by the character of the load into which the antenna operates. A foreign body in the ground under the antenna changes the oscillating conditions and affects the bias on the low-frequency amplifier so as to increase the output of the 1000-cycle wave derived from the 1N5 oscillator.

2.3 Electromagnetic Methods

2.3.1 Theory

Electromagnetic methods apply only to mines having metallic parts. If a magnetic object is brought close to a coil, the self-inductance of the coil increases. If the object is electrically conductive, the self-inductance decreases. In both cases, if the object is small and if its distance from the coil is large compared with the radius of the coil, the variation of self-inductance is inversely proportional to the sixth power of the distance between the coil and the object.

It may also be noted that if two coils are used, and if the object is brought close to the assembly of coils, the mutual inductance between the two coils also varies. If one coil is supplied with alternating current, the voltage detected at the terminals of the second coil varies with the nearness of the object. The variation is directly proportional to the frequency of the current if the object is only magnetic, and proportional to the square of the frequency if it is only a conductor. In the latter case, the voltage induced by the approaching object is not in phase with that obtained in the absence of the object. The phase angle depends on the electrical resistance offered by the object to the currents developed, the resistance of the receiving coil, and the self-inductance of these two elements.

An object at a certain distance from the coil system can have such magnetic and conducting properties at a single frequency as to produce no noticeable effect. At a higher frequency, the object will behave like a conductor and, at a lower frequency, as a magnetic substance.

The two coils may be adjusted to have zero mutual inductance when far from any metallic object, and to generate an alternating electromotive force in the receiving coil when such an object is approached.

2.3.2 Variation of Self-Inductance

If a single coil forms the inductance of a resonant circuit maintained in oscillation by a tube, the self-inductance variations will cause corresponding changes in the frequency of oscillation. If this frequency is fairly high, and if beats are detected between this oscillator and a second oscillator of a fixed frequency fairly close to the first one, the resultant beat note changes appreciably in pitch when approaching a metallic object.

This arrangement was used in the S.F.R.441, a French apparatus that is now obsolete. At the time it was designed, it detected satisfactorily the large metal mines then exclusively used.

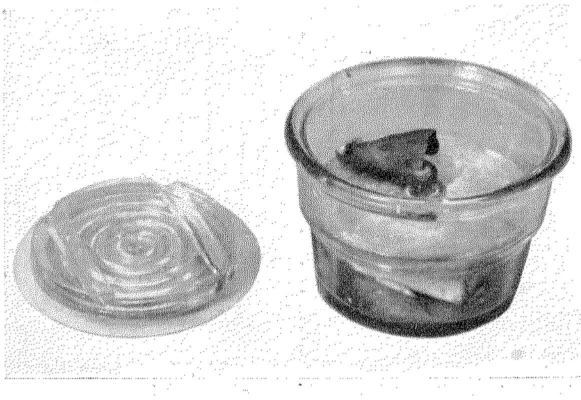
2.3.3 Variation of Mutual Inductance

2.3.3.1 Feedback Apparatus

The simplest type of feedback or "singing" equipment consists of an oscillator in which the coupling is slightly below the value necessary for sustained oscillations. If, at the frequency to which the circuit is tuned, the object to be detected has very low resistance and is strongly magnetic, there will be self-oscillation over a range where the variation in mutual inductance is positive and maximum; the variation in self-inductance is always positive in this case. If the object is only conductive, oscillation will occur over another range of inductance. If the object is a poor conductor, there is a decrease in the spacial region over which oscillation occurs and, generally, a decrease in sensitivity.

In addition, oscillation frequencies differ for a magnetic object and a conductor; the frequency of oscillation for a magnetic body being lower than that for a conductor. In practice, the performance of the detector is influenced also by the shape of the object.

A magnetic object may act normally at a certain distance from the detector and behave



A glass antipersonnel mine, with cover removed. It was generally provided with a steel plate on which a lever-type fuse is mounted.

like a conductor at another distance, depending on the manner in which the lines of force are intersected.

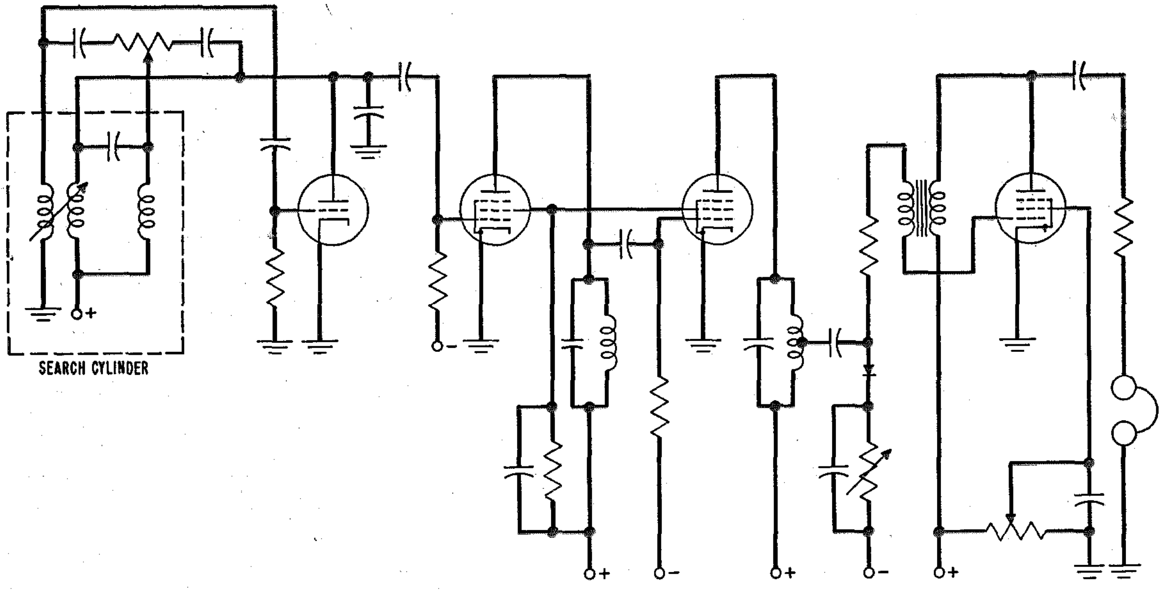


Fig. 2—Circuit arrangement of the German Wien 41 detector. Proximity to a mine sets the first high-frequency oscillator in operation. The two pentode amplifying stages and detector produce a biasing voltage which initiates oscillations in the final audio-frequency circuit.

Finally, the oscillating system is seldom simple. It is frequently an amplifier with several selective stages and employs a feedback circuit. The main difficulty with apparatus using feedback is the risk of getting further away from the condition of sustained oscillation without its being noticed, and thus causing a decrease in sensitivity. This may result from a variation in supply voltage which affects the amplification and internal resistance of the tubes. It is necessary, therefore, to provide very stable supply voltages.

2.3.3.2 British Detector No. 4

The British No. 4 detector consists of an audio-frequency amplifying system, carried on the back of the operator, a feedback adjustment unit on the shoulder, a search plate containing the coupling device at the end of a telescopic handle, and headphones. The design is good from the point of view of sturdiness, waterproofing (which is complete even to the headphones), stability, and size of the search plate. The only criticisms that could be made are the absence of a device to check the operation independent of the adjustment knob, and difficulty in the preliminary adjustment of the search plate. Its sensitivity is satisfactory, as the fuse of the *Schümine* may be detected.

2.3.3.3 German Detector Wien 41

The general appearance of the Wien 41 detector is similar to that of the British No. 4 type, but the search system is at the end of a tube, a molded cylinder which can be moved with its axis parallel to the ground. The circuit is shown in Fig. 2. A high-frequency oscillator is adjusted slightly below the oscillating point. It is followed by a two-stage pentode amplifier and a rectifier. The direct voltage thus obtained when the oscillator is working is used to operate a second oscillator tuned to an audio frequency and also maintained normally below the singing point. It is, therefore, a double-threshold instrument. Power is obtained from a self-rectifying vibrator and a small storage battery. The shape of the instrument is convenient, but its sensitivity is slightly lower than that of the British design.

2.3.3.4 Balanced-Type Instruments

In the balanced-type instrument, an audio-frequency oscillator is in constant operation. In the search cylinder, the plate inductor of the oscillator is oriented to have minimum coupling to a tuned receiving circuit connected to the input of an audio-frequency amplifier. The presence of a mine will distort the field from the oscillator coil and produce a signal in the receiving circuit. The output of the amplifier may be connected to a headphone or a meter. The variation in self-inductance and the frequency shift of the oscillator caused by the proximity of the mine can usually be neglected. Electrical balancing elements are normally required to compensate for the capacitive unbalances and undesirable couplings caused by the physical arrangement of the components and wiring.

2.3.3.5 German Detector Frankfurt 42

A symmetrical oscillator using two triodes provides the operating signal in the Frankfurt 42 mine detector. The circuit is given in Fig. 3. Two selective pentode audio-frequency amplifying stages and headphones produce a usable signal from the unbalance voltage picked up by the receiving coil. Balance is ensured both by an adjustment of the search element and by the injection, at the first amplifying tube, of a suitable voltage from the differential capacitors *C1* and *C2*. The power supply is obtained from a storage battery and a vibrator.

The mechanical arrangement resembles that of the Wien 41; the electrical apparatus is carried in a pack on the back of the operator. The search cylinder is at the end of an adjustable tubular handle. The sensitivity of this instrument is greater than that of the Wien 41.

2.3.3.6 French Detector S.F.R.451 or DM-2

The arrangement of the Wien 41 detector was adapted to a French design known as the S.F.R.-451 or DM-2. An illustration of the instrument in operation is given.

The S.F.R.451 or DM-2 consists of a search element of tubular shape, carrying two knobs for adjusting to a minimum the voltage induced by

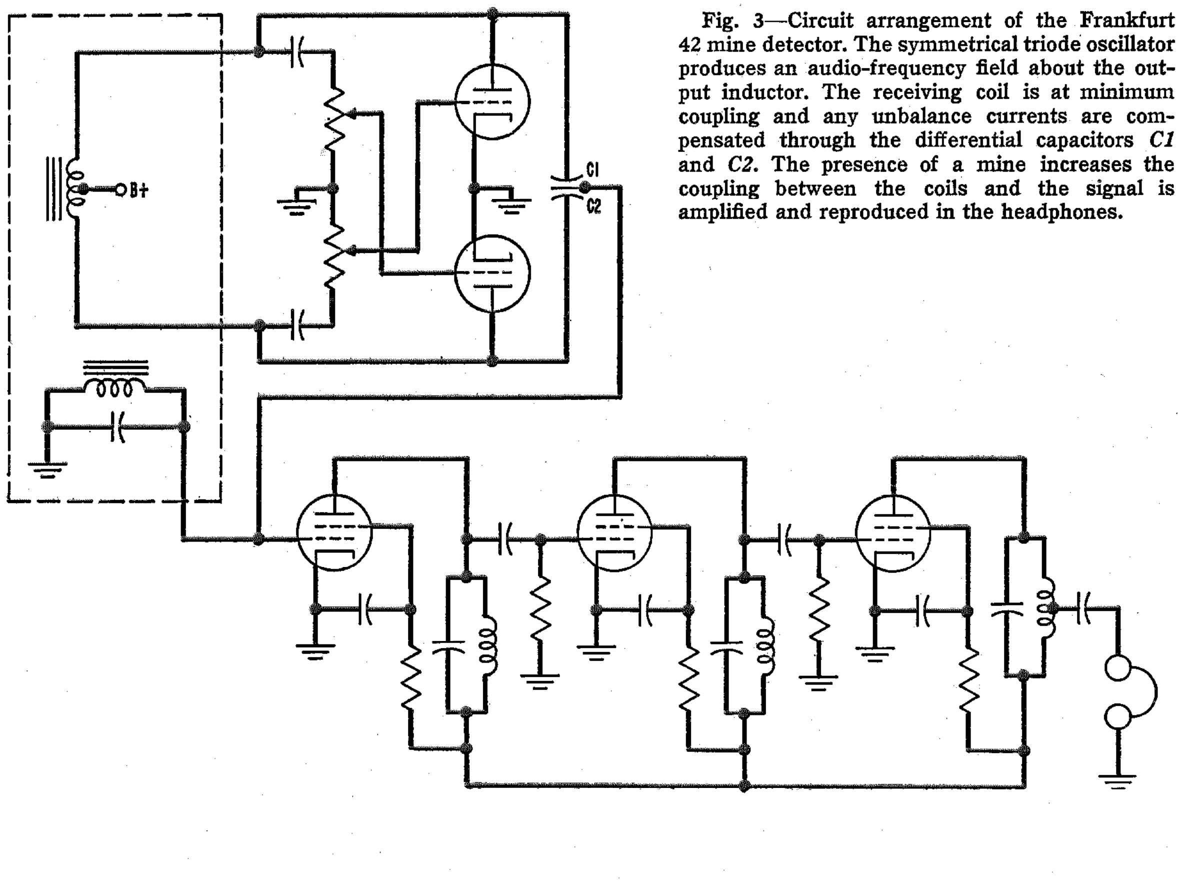


Fig. 3—Circuit arrangement of the Frankfurt 42 mine detector. The symmetrical triode oscillator produces an audio-frequency field about the output inductor. The receiving coil is at minimum coupling and any unbalance currents are compensated through the differential capacitors *C1* and *C2*. The presence of a mine increases the coupling between the coils and the signal is amplified and reproduced in the headphones.

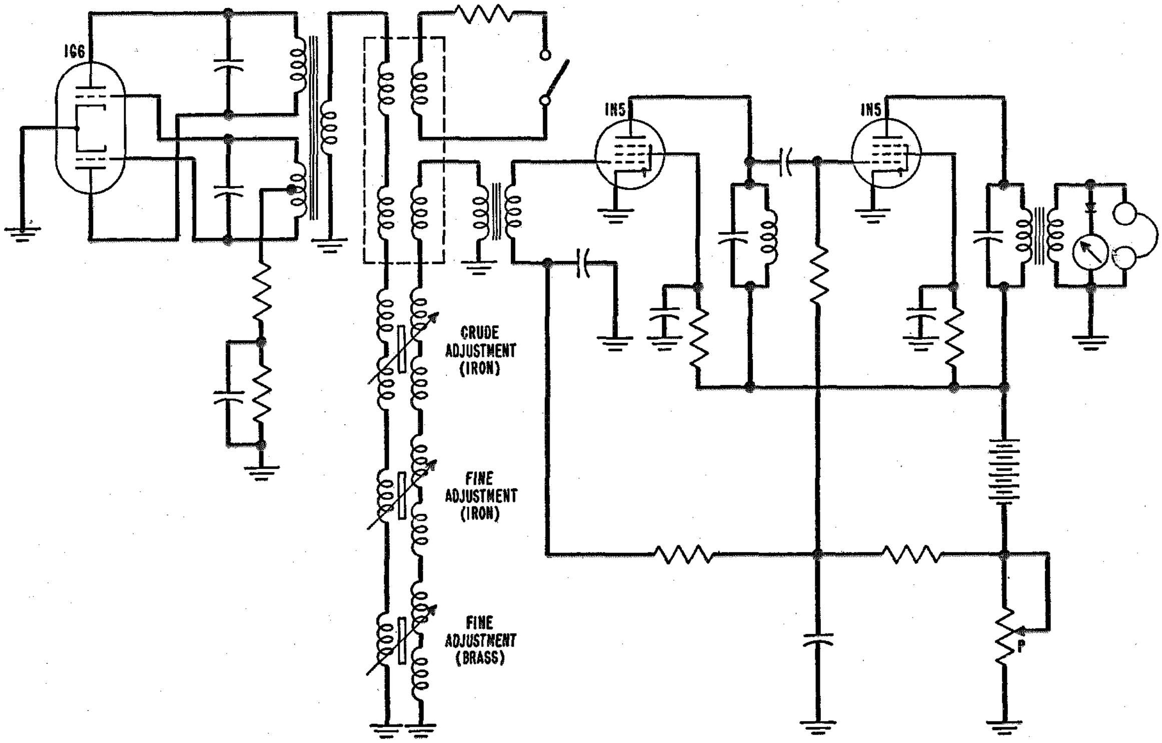
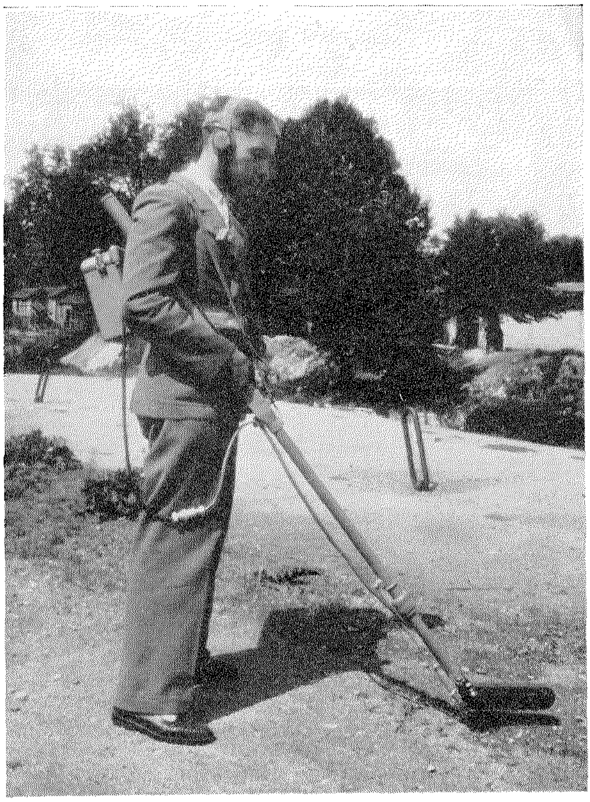


Fig. 4—Circuit arrangement of the American detector SCR-625. The 1G6 produces 1000-cycle oscillations which pass through a step-down transformer to the search coils. Compensating voltage for balance is controlled in amplitude and phase by the three adjustment transformers. A two-stage tuned amplifier operates a loudspeaker and rectifier-meter. The extra coil in the search system is for testing and produces a known deflection of the output meter when its switch is closed.

the oscillator in the receiving coil. This search element may be connected to an adjustable handle that is also collapsible for ease in transport. The tubes, circuits, and dry batteries are housed in a box, which may be carried on the back of the operator in a manner similar to Alpine knapsacks.

The sensitivity control may be fastened either to the strap of the box or to the operator's belt. With correct adjustment of the instrument, a harmonic of the oscillator current is heard in the headphones. This tone is different from that caused by the proximity of a mine; consequently, it allows the operation of the equipment to be monitored continuously.



French detector type S.F.R.451 or DM-2 adapted from the German Frankfurt 42, which it closely resembles in appearance. This device may be used with the operator in a prone position by adjusting the handle to be horizontal.

2.3.3.7 American Detector SCR-625

A 1000-cycle oscillator, consisting of a 1G6 double triode, energizes the transmitting coil through a step-down transformer, as may be seen in Fig. 4. This inductor is in two sections and is coupled to a single coaxial receiving coil. The compensating voltage is injected in series with the receiving coil through three small differential transformers with adjustable cores. A main iron-core transformer is designed for preliminary adjustment at the factory. Two fine-adjustment transformers, one having an iron and the other a brass core, allow an accurate balance to be made by injecting a voltage which may be controlled both in amplitude and in phase, respectively.

The receiving coil is coupled through a high-ratio step-up transformer to a pentode with a tuned plate circuit followed by a second pentode, the output transformer of which is also tuned. The amplified signal operates a loudspeaker and a rectifier-voltmeter. The amplifier gain is adjusted by a potentiometer controlling the bias of the two pentodes.

The short-circuiting of a turn in the search plate unbalances the system, the amplifier gain being adjusted so that this condition corresponds to a calibrated deflection of the visual indicator.

The power supply is obtained from dry batteries. The filament of the oscillator tube is supplied from a separate battery to avoid coupling to the amplifier. When the battery voltage varies, the frequency of the oscillator also shifts, and it is necessary to change the adjustment of the compensators, the operation of which is correct only for a single frequency.

A sheet-iron box, carried in a pack slung across the shoulder, contains the oscillator, amplifier, gain-control potentiometer, and batteries. The loudspeaker is fastened by a strap to a belt over the shoulder of the operator.

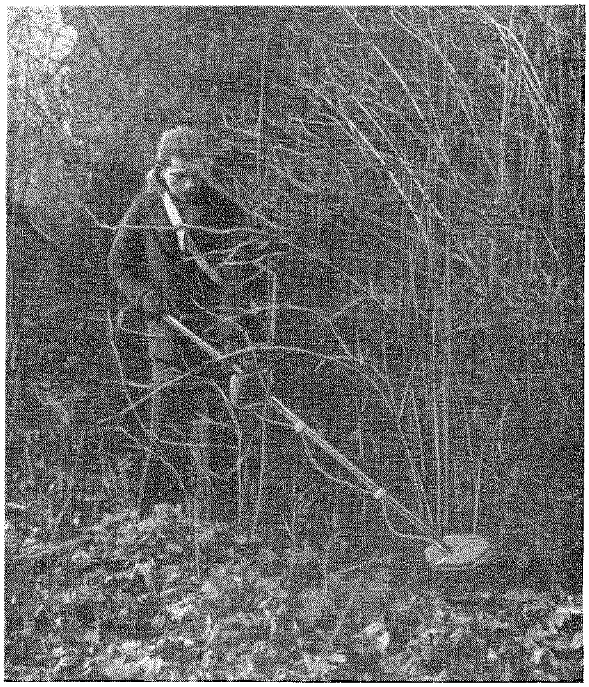
The search coil assembly, terminated by the search plate, consists of a control unit containing the balancing transformers, control switch for the test loop in the search plate, main switch, and visual indicator.

The search plate is constructed of light wood covered with waterproof cloth. Maximum sensitivity is obtained when the object is on the plate's axis. The graphical representation of the induced voltage is somewhat similar to a highly damped resonance curve, sensitivity decreasing rapidly with distance.

The rectifier-voltmeter indications are nonlinear. For one unit of input, a meter deflection of approximately 60 percent of full scale occurs. Full-scale indication, however, requires 10 relative units of input. This provides relatively greater sensitivity at low signal levels. At the same time, the total deflection is arranged to correspond to the maximum voltage obtained from the amplifier in normal operation, thus eliminating any overload of the instrument. The main figures on the scale and the tip of the pointer are covered with a phosphorescent paint so as to be visible at night.

The loudspeaker diaphragm and cavity have resonant frequencies that produce an acoustical response curve having the same appearance as that of a band filter. The unit provides high attenuation of the second harmonic and gives almost constant output over the pass band. This is an appreciable advantage from the manufacturing standpoint.

The SCR-625 detector is extremely sensitive and very reliable in use. The instrument has been manufactured in very large quantities in the



Mine detector DM-3 in operation over difficult terrain.

United States, and was used by the American army in almost all of its operations.

3. Le Matériel Téléphonique

3.1 SCR-625-FR

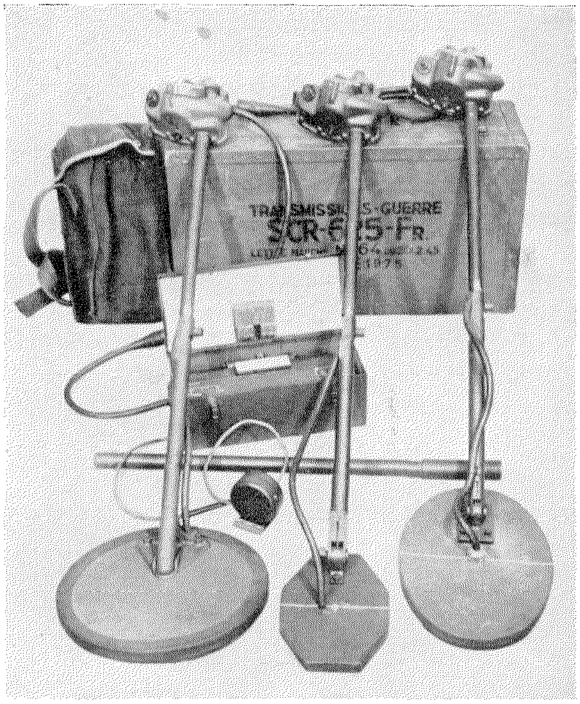
The instrument just described has also been manufactured for the French government as the SCR-625-FR, shown in an accompanying illustration. The Boulogne works of Le Matériel Téléphonique produced these equipments at the rate of 2000 per month. Special studies were made to adapt the original design to the French raw-material supply conditions.

3.2 DM-3 AND DM-4 INSTRUMENTS

Despite its high sensitivity, and partly because of it, the SCR-625 has, in some cases, the disadvantage called "ground effect." When the distance between the search plate and the ground is varied, the flux may change, producing an effect similar to that when a metal object is present. Therefore, the plate must be held about 8 inches above the ground to minimize accidental variations in relative position. The magnitude of this effect depends on the nature of the ground. Road surfaces containing iron slag from blast furnaces are magnetic and cause a positive variation in mutual inductance; others may contain iron and manganese salts and are paramagnetic, giving an effect similar to that of magnetic soils, but much weaker for an equal concentration. Finally, others are conducting and give an opposite but, in general, a still-weaker effect.

The ground effect, except in the case of definitely magnetic surfaces which are fairly rare, is important only because of the large quantity of matter covered by the lines of force from the transmitting and receiving systems. One solution would involve a special design of coils to produce fields of such shape that the volume of active soil would be small. The ideal solution would be an alternating magnetic field in the shape of a "hair pin," reaching deep into the ground with high intensity and forming a narrow tube of force. Such an arrangement would involve very complex transmitting devices.

A simple solution has been found by engineers of Le Matériel Téléphonique. It is based on the assumption that the soil is homogeneous and the magnetic density and resistivity are constant and diffused over large volumes, while those of the mines are not diffused.



SCR-625-FR type equipments were manufactured in France by Le Matériel Téléphonique in substantial quantities. Based on the American design, they were modified to meet the local raw-materials conditions. The DM-3 and DM-4 instruments, having octagonal and elliptical search plates, respectively, are also shown.

Fig. 5 shows the principle of the search plate. The transmitting coil E has an axis of symmetry Y, while the receiving system consists of two identical coils $R1$ and $R2$, symmetrical with respect to the same axis.

If the system is kept parallel to the surface of a homogeneous volume of any nature, the effects in the two receiving coils cancel each other. If, however, a magnetic or conducting object is moved at a constant height from left to right in front of the system, it acts, first, more on the mutual inductance of E with respect to $R1$ than on that of E with $R2$, and the resultant voltage goes through a maximum.

When the object passes the Y axis, the two voltages cancel exactly and, beyond that point, the previous effect is reproduced with greater effect on $R2$. The voltage minimum is in the form of a cusp and is very sharp. It locates exactly the position of the object. On the other hand, a lateral displacement across the X axis yields the usual response curve.

In the first tests, the handle was oriented along the Y axis. The result, in the usual sweep motion, was two maxima separated by a minimum for each metal object and an accurate localization was fairly delicate. This pattern of sounds confused the mine-disposal operators who were accustomed to working with only a minimum.

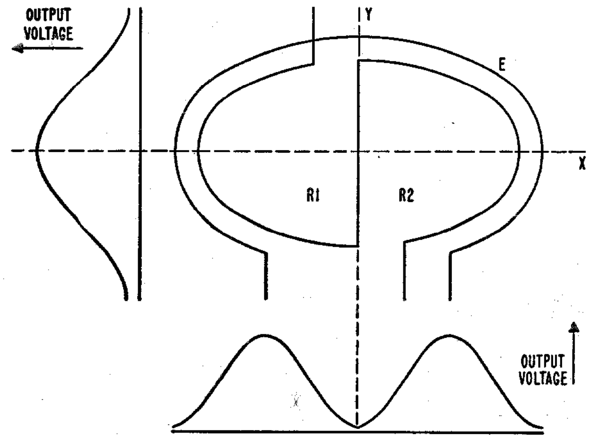


Fig. 5—Search plate of DM-3 and DM-4 units. The fields from *E* develop in *R1* and *R2* equal and opposite voltages which balance in the absence of a foreign body. The response to moving a magnetic or conductive object near the coils is plotted for two directions of motion in the small adjacent graphs.

Placing the handle along the X axis avoids these drawbacks; detection first gives a single maximum in the sweep motion, then the operator hunts for the point of extinction by swinging back and forth.

Besides eliminating the ground effect, this search system permits working in the vinicity of telephone lines and power lines in cables, because nearly uniform extraneous magnetic fields produce equal and opposite induction effects. For the same reason, two instruments may be operated close to each other. The increased accuracy in mine localization permits separate detection of mines close to each other, even those of widely different types.

An instrument based on the same principles has been designed for mine detection along railroads and in the vicinity of other large metallic masses. If the SCR-625 detector and a compensated-plate detector have the same sensitivity for a large object at a certain distance, the compensated-plate detector is more sensitive to a small object at a shorter distance. This gives better results in hunting for small antipersonnel mines.

The DM-3 instrument is based on the above principle and comprises a small octagonal rotating search plate, especially designed for irregular surfaces, such as gutters and bushes. Its small size results in an appreciable reduction in possible sensitivity.

The DM-4 instrument differs from the DM-3 only in that the plate dimensions are larger and the plate is elliptical. Its sensitivity is equal to that of the SCR-625 instrument for detecting the fuse of the *Schümine,* and is slightly less for large mines. It can be used on the ground surface and its effective sensitivity is definitely greater than that of the SCR-625, especially for small mines.

These two types have been manufactured on a production basis at the Boulogne works of Le Matériel Téléphonique and, together with the SCR-625 equipment, the number made has reached 16,000. This constitutes the majority of the mine detectors used in France. They sometimes have unexpected applications, and have been employed for the detection of metal splinters in tree trunks to avoid damaging the sawmill and for the location of underground cables, documents, or buried valuables.

4. Conclusion

In a country devastated by enemy occupation and war, the rapid manufacure of mine detectors, undertaken only a few months after the liberation of Paris, seemed an impossible task. It was accomplished, nevertheless, despite endless supply difficulties. In particular, thanks to the United States Army, most of the apparatus was equipped with tubes and batteries taken from allied stocks. By way of appreciation for its contributions to the war effort, Le Matériel Téléphonique received the coveted "A" award of the U.S. Army.

At the present time, the removal of mines is completed, and all the mined areas, including agricultural lands, beaches, forests, etc., have been restored to public use with a minimum loss of human lives.

Speech-Input Equipment for New Oslo Broadcasting House

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ERECTION of the Norwegian Broadcasting House in Oslo was started before the recent war, and the building itself, without interior equipment, was half finished when the country was occupied. The speech-input equipment, comprising the switching system, control desks, microphone and main amplifiers, etc., was not installed until nearly two years after liberation. The recording and reproducing apparatus is partly new and partly older equipment.

• • •

1. *General*

Most of the studios, of which there are twenty in all, are located on the ground floor of the Norwegian Broadcasting House, as may be seen in Fig. 1. They are arranged in a separate block consisting of four main groups.

A. The great concert studio (2), with an announcing studio (1).
B. 4 music studios (8, 9, 10, 11) and 1 speech studio (7).
C. 4 speech studios (3, 4, 5, 6) for lectures and interviews.
D. 4 entertainment and dramatic studios (12, 13, 14, 15).

Elsewhere in the building, there are 5 additional studios for news and rehearsals.

As will be seen, the studios of groups *A*, *B*, and *D* are visible from control rooms. There are four 4-channel (*SK*) desks and two 8-channel (*RK*) desks. There are also four control rooms having no visual access to studios; two of these are main control rooms, one is a recording control room, and the other is a special control desk located in the amplifier room. All music and dramatic studios are normally operated from the neighboring *SK* and *RK* rooms, whereas programs from the speech studios may be controlled from either an *SK* or an *RK* room, or directly from a main control room.

Additional program sources are the 10 reproducer rooms, of which three contain recording apparatus. Other technical rooms for the program chain include seven recording rooms, one amplifier room, one relay switching room, one studio private automatic branch exchange and rectifier room, and one battery room.

The dominant feature of this equipment is that completely automatic switching has been introduced, both for the selection of program sources to the control-desk input channels, and for the output of the desks to the various transmitters and recorders. Automatic switching is employed not only for program lines but also for all of the signaling equipment, which at this installation is rather comprehensive. It would have been too complicated and inconvenient to use manual switching and, undoubtedly, fewer errors occur in automatic operation. The automatic system includes relays and finders similar to those that have amply proven their reliability of operation in automatic telephone exchanges.

2. *Speech-Input Equipment*

Fig. 2 shows a block diagram of the program chain and a corresponding level diagram. Zero level in decibels on the outgoing line corresponds to 1 milliwatt in 600 ohms, i.e., 0.775 volt, which was established as the reference level for audio-frequency measurements at an international conference just before the war. A maximum level of +6 decibels (1.55 volts) is permitted. The output from the program source is amplified in a microphone amplifier to a level at which switching and adjustment may be made with a minimum risk of disturbance. After the mixer, a main amplifier raises the level to the maximum permissible output and the outgoing program is distributed from a program bus. Each outgoing line incorporates a line or isolation amplifier, which provides no gain and serves only to prevent disturbances from reaching other lines connected to the same program bus. The inputs of loudspeaker and volume-indicator amplifiers are also connected to the program bus.

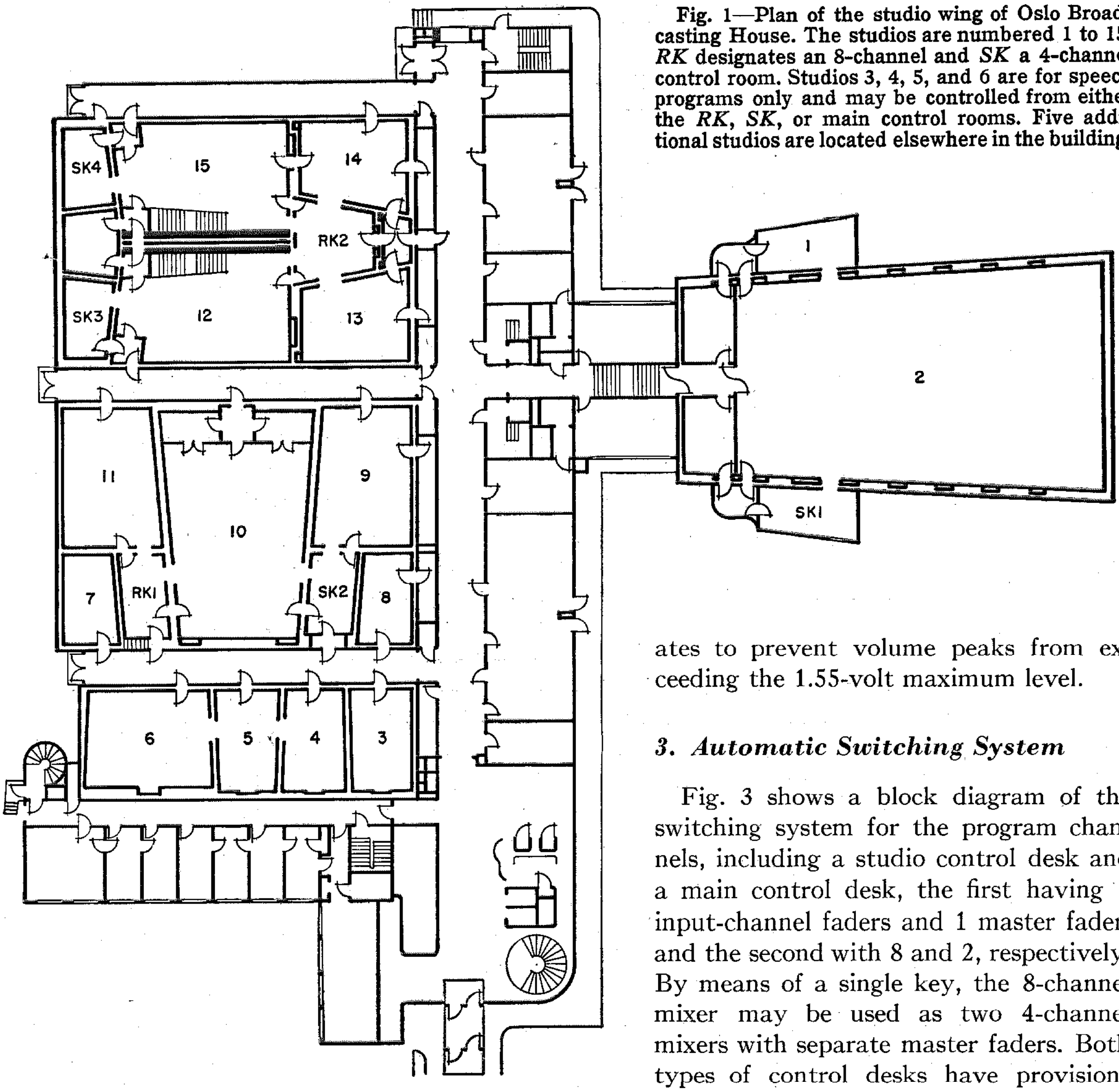


Fig. 1—Plan of the studio wing of Oslo Broadcasting House. The studios are numbered 1 to 15. *RK* designates an 8-channel and *SK* a 4-channel control room. Studios 3, 4, 5, and 6 are for speech programs only and may be controlled from either the *RK*, *SK*, or main control rooms. Five additional studios are located elsewhere in the building.

The total maximum gain is about 110 decibels. Usually, the microphone amplifier has a gain of 35 to 50 decibels, depending on the type of program and the microphone used. The output level of the microphone amplifier is always adjusted to about −24 decibels, this being the input level to the control desk. The insertion loss of the mixer is 10 to 13 decibels, plus an additional attenuation of about 10 decibels to allow for raising low levels of musical programs, and to allow for unusually high attenuation in program lines. The gain of the main amplifier is adjusted to 50 decibels. Incorporated in the main amplifier is a limiting device, which operates to prevent volume peaks from exceeding the 1.55-volt maximum level.

3. *Automatic Switching System*

Fig. 3 shows a block diagram of the switching system for the program channels, including a studio control desk and a main control desk, the first having 4 input-channel faders and 1 master fader, and the second with 8 and 2, respectively. By means of a single key, the 8-channel mixer may be used as two 4-channel mixers with separate master faders. Both types of control desks have provisions for selecting a certain number of a total of 93 program sources, such as microphones, reproducing apparatus, and incoming program lines. The main-control-room desks may, in addition, select the output channels of the various studio desks as program sources.

Seven of the 12 outgoing lines or program buses connect to recording apparatus, and any one of these may be switched to the output channel of a control desk.

3.1 CONTROL SIGNALING

3.1.1 Lamp Signals

Both white and red lamps are mounted inside and outside of the studios, and also on lamp

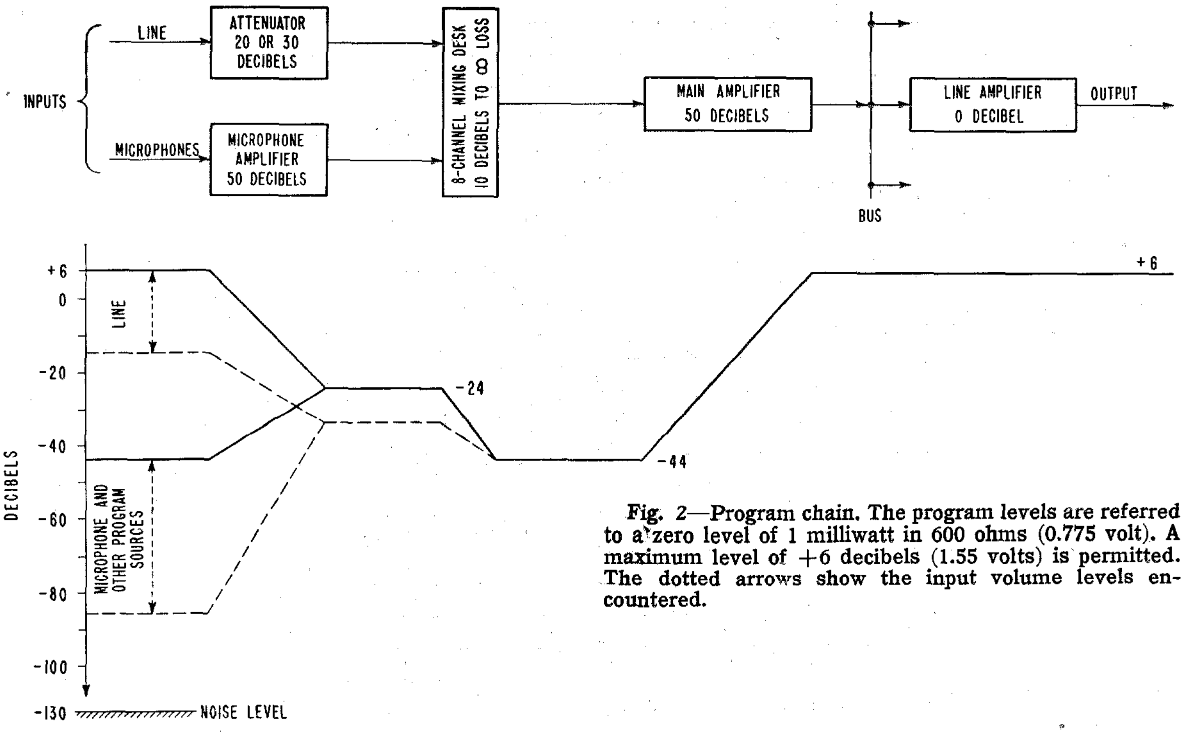
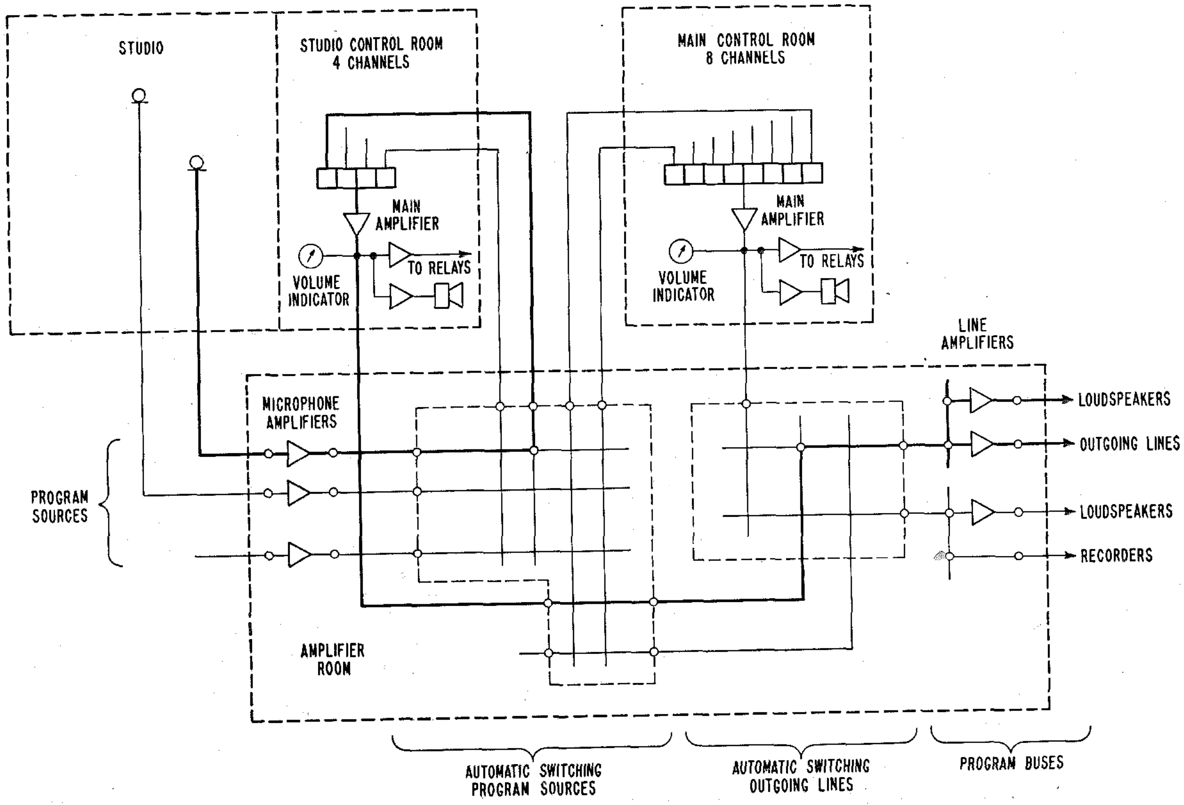

INPUTS
LINE
MICROPHONES
ATTENUATOR 20 OR 30 DECIBELS
MICROPHONE AMPLIFIER 50 DECIBELS
8-CHANNEL MIXING DESK 10 DECIBELS TO ∞ LOSS
MAIN AMPLIFIER 50 DECIBELS
BUS
LINE AMPLIFIER 0 DECIBEL
OUTPUT
DECIBELS
+6
0
-20
-40
-60
-80
-100
-130
LINE
MICROPHONE AND OTHER PROGRAM SOURCES
-24
-44
+6
NOISE LEVEL

Fig. 2—Program chain. The program levels are referred to a zero level of 1 milliwatt in 600 ohms (0.775 volt). A maximum level of +6 decibels (1.55 volts) is permitted. The dotted arrows show the input volume levels encountered.

Fig. 3—Block diagram of switching system for program channels.



panels at various locations in the building. The white lamps are lighted automatically when the studio has been connected to a control desk, and are replaced by red signals when the control-desk fader is opened. Also, a red lamp at the microphone in use (there may be 2 to 6 microphones per studio) will light up.

3.1.2 Acoustic Signals

When a studio has been connected to a desk, any program then passing through the selecting control desk will be reproduced by a loudspeaker in the studio, enabling the persons present to follow the program. The loudspeaker is automatically disconnected when the red light appears. The actors may then listen to the program with headphones, which may be necessary, for instance, when two or more studios take part in the program. Practice has shown that the red light is not sufficient for attracting the attention of the actors, and a warning note is therefore given through the loudspeaker by pressing a push-button on the control desk before the fader is advanced and the red light appears. In the music studios, the loudspeaker may also be connected to an oscillator giving the concert pitch for tuning the instruments.

3.1.3 Talk-Back System

The studio loudspeaker may also be automatically connected to a talk-back microphone on the control desk without disconnecting the studio microphones, thereby establishing a duplex connection between studio and control room. This system is especially of value when rehearsing radio plays.

3.1.4 Studio Signal

Each studio contains an announcer's set for communication with the control operator. On pushing a button on this set, a green lamp at the control-desk input channel to which the microphone is connected will light until the position

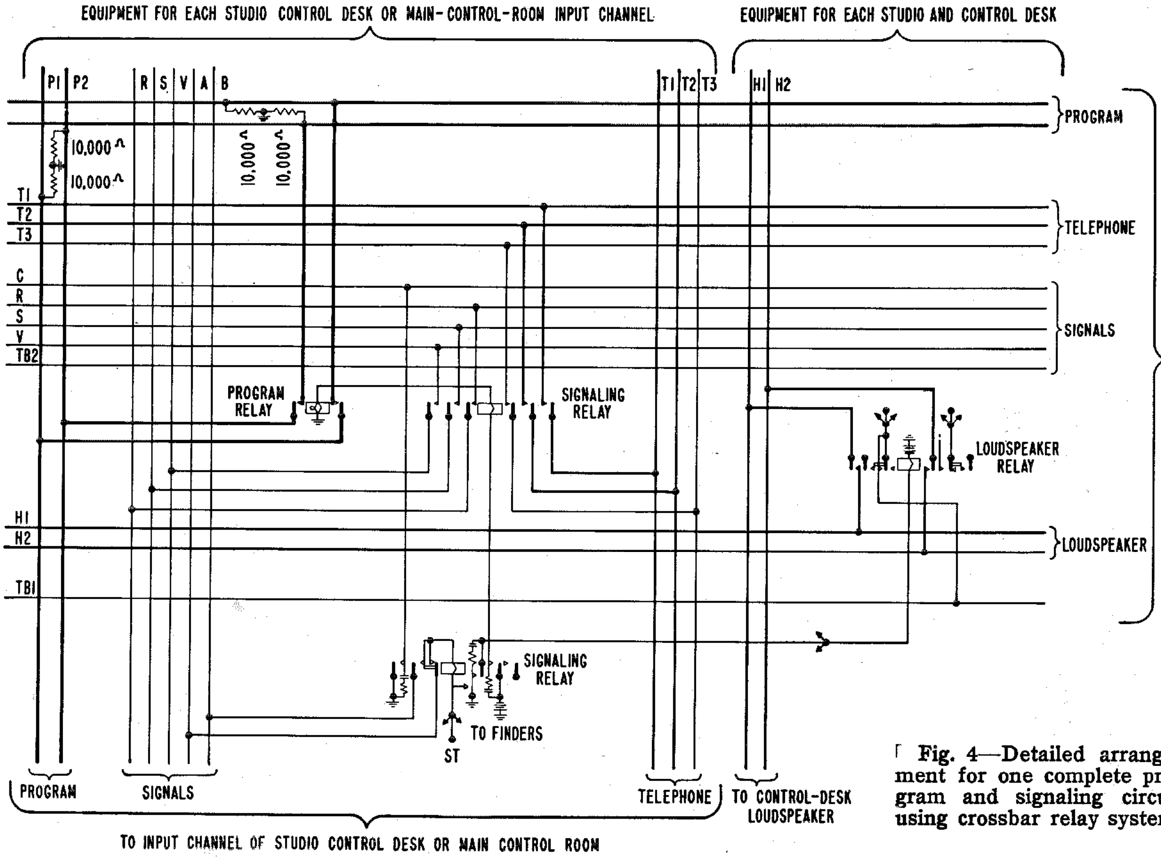


Fig. 4—Detailed arrangement for one complete program and signaling circuit using crossbar relay system.

of the input fader has been changed from closed to open, or vice-versa. This signal is used by the announcer when requesting a microphone and after finishing an announcement.

3.1.5 Telephone

The announcer's set includes a telephone which, when a studio is not selected, is connected to a private automatic branch exchange serving all studios and technical rooms. When selecting a studio, the telephone is automatically switched to the control room in question for direct communication between announcer and operator.

3.2 SWITCHING EQUIPMENT

Automatic switching is accomplished by a relay crossbar system. Fig. 4 is a detailed schematic of the connections of the program sources and signaling lines between studios, studio control desks, and main control rooms; this corresponds to any one of the crossing points in the switching of program sources illustrated in Fig. 3. For each microphone and each input channel, one low-capacitance relay (program relay) connects the program line, and over the contacts of two other relays (signaling relays) the signaling lines are connected. In addition, there are one or two relays for each studio and control room connecting the loudspeaker lines. The operation of these relays is controlled through 9-by-100-point finders from a control circuit, which receives impulses from a keyboard on the control desk. Similar circuits with minor deviations exist for connecting reproducers, outside program sources, and outputs of control desks to input channels. The automatic switching system for the selection of recorders and outgoing lines is also built up in the same manner.

It should be noted that neither the program nor the signaling currents pass through the finder contacts, but rather through the contacts of the relays at the crossing point.

For the selection of program sources, four digits are sent. The first indicates the number of the input channel to which the program source is to be connected, the two following give the number of the studio, and the last one is the number of the microphone in the studio. The outgoing-program lines and recorders are selected similarly with the exception that only 3 digits are sent, the first being the number 9, indicating the output channel, and the two following corresponding to the number of the program bus. For each input and output channel of the control desk, there is an indicator which, when a connection has been established, shows the number of the program source or the bus selected.

Theoretically, there are more than 5000 possible connections between the program and signaling lines. Only about 1200 are equipped, but facilities are available for extension to about 2200. The layout of the relay system, including program and signaling relays, is such that the relay bays actually provide space for all of the 5000 possible connections. Cabling, however, is provided only for the connections that are expected to be used, 2200 in all. This arrangement provides great flexibility. Changes and extensions

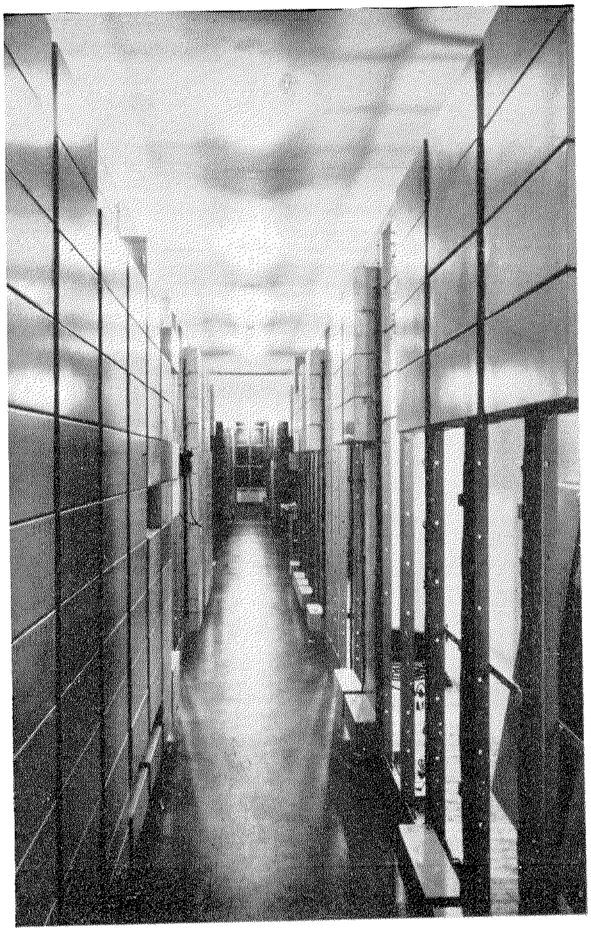


Fig. 5—Relay room.

may easily be made by mounting the necessary relays at the corresponding crossing points without disturbing the operation of the equipment already in use. The program and signaling relays are mounted on separate bays. There are 22 program-relay bays and 36 signaling-relay bays, including the output-switching system.

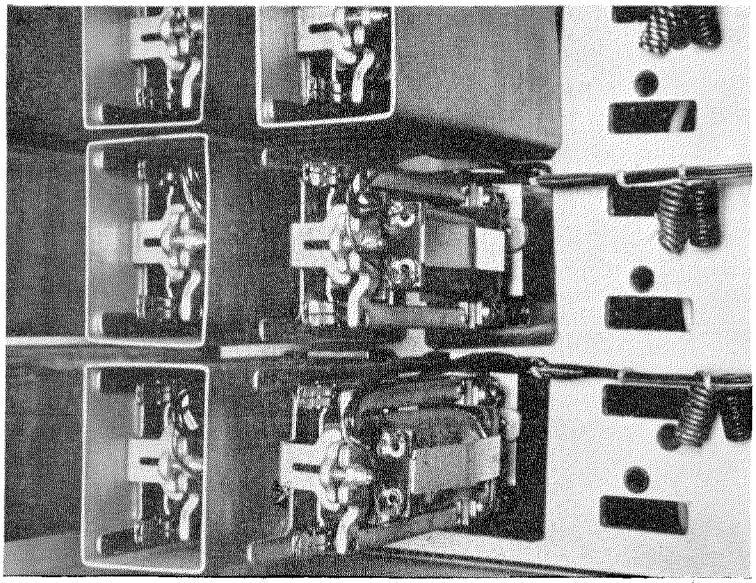


Fig. 6—Front view of program relays as mounted. Relays are omitted for positions not considered necessary for normal operation. They may be installed at any future time if they are needed.

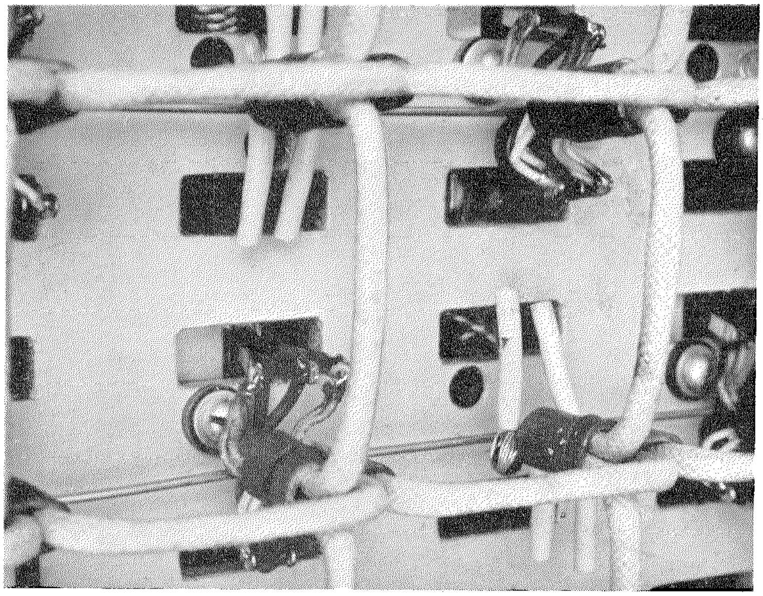


Fig. 7—Rear view of program-relay panel. The vertical wires are connected to program sources and the horizontal wires go to the various input desks. To avoid cross talk and noise, the wires are shielded and covered with an insulating braid.

Fig. 5 is a view of the relay room. The rows in the foreground contain the program relays; those in the background the signaling relays. In this part of the room are also located 9 bays for such miscellaneous circuits as lamp signals, acoustic signals, telephone systems, amplifier-switching circuits, and alarm circuits, in addition to a control unit, a routine test circuit, and 3 finder bays.

Emergency patching of the connections is possible in case of breakdown of the automatic system. Only the most important signaling connections may be established in this way, however, to avoid complexity of operation under emergency conditions.

With such an extensive automatic system, it has been necessary to protect program lines from clicks and other disturbances caused by the rapid charge and discharge currents that may occur when two program lines having different induced potentials are connected together. Each section of wire between amplifiers is grounded through 10,000-ohm resistors as shown in Fig. 4. A 0.1-microfarad capacitor in series with 50 ohms is connected across signaling-relay contacts to quench any sparking. All relays are equipped with bifurcated contacts to reduce failures from bad contacts.

The requirements as to noise are very stringent. The noise level at the input of the microphone amplifier must not exceed 0.25 microvolt when referred to a source impedance of 50 ohms. This corresponds to an absolute noise level of 130 decibels below zero level (1 milliwatt in 600 ohms). The cross-talk voltage induced in one program channel from another must not exceed the noise level just specified.

Measurements prove that these requirements have been met.

Some notes on the steps that have been taken for obtaining high-quality transmission may be of interest: All program wires are screened, either with lead sleeve or copper screen. The screen is covered with a braiding of insulation to prevent it from touching any metallic parts such as cable racks, bays, etc. All screens are connected to a ground terminal in the amplifier room and from there to a separate ground plate. Care is taken that no induction loops are formed in the screening.

As previously mentioned, the program relays are mounted separately from the signaling relays. Each program relay is equipped with a cross-talk cover. Further, a grounded spring is located between each set of relay springs to eliminate possible cross talk from other program sources. The relays are mounted on plates of nonmagnetic material. The program wires are connected to the relay springs at the back of these plates, whereas the coil connections are made on the front and are so connected that the outer layer of the winding is at ground potential.

Figs. 6 and 7 show the front and back of a program-relay mounting plate. The vertical wires represent the program sources, and the horizontal wires are for the desk input channels.

4. Control Desks

There are three types of control desks: 4-channel desks, 8-channel desks as shown in Fig. 10, and a special desk located in the amplifier room. This special desk includes, in addition to the equipment for a 4-channel desk, the necessary apparatus for supervising and connecting all incoming and outgoing lines. It differs, accordingly, in many respects from the others.

As in the automatic switching system, great attention has been paid in the desks to the elimination of possible interference from the signaling lines to the program channels. The apparatus for signaling has been separated as

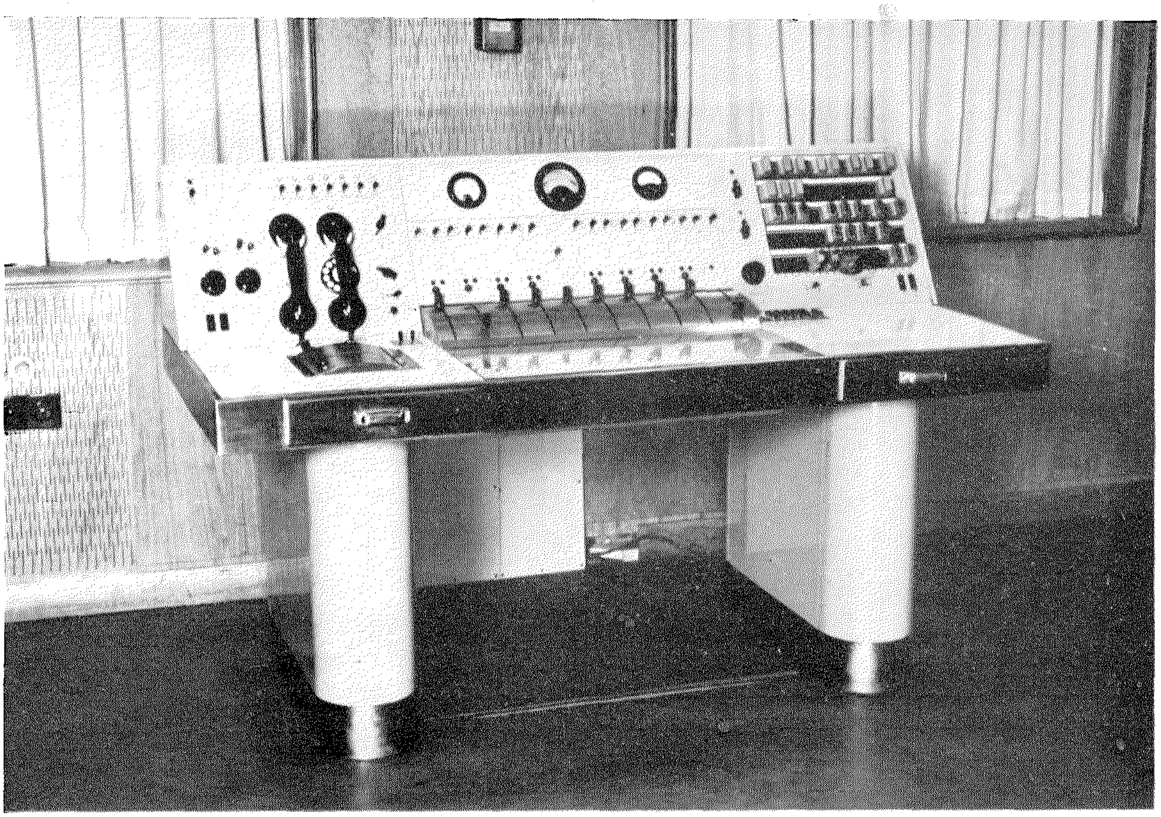


Fig. 8—8-channel control desk. All program circuits terminate in one pedestal and the signaling circuits terminate in the other. Ten faders are mounted at the center and provide for the 8 channels and the two master controls. A jack panel for setting up special connections is located at the right.

far as possible from the program equipment. The program wires are connected to terminal strips in one pedestal of the desk whereas all of the signaling cables are terminated in the other pedestal.

It is beyond the scope of this article to describe in detail all the apparatus in these desks. It may be noted, however, that on one of the panels is a main switch, operated when the control desk is placed in use, to connect power to the desk and amplifiers in the control room and to operate the warning lamps indicating that the control room has been engaged. The faders are of the sliding type, with handles movable toward the operator; the maximum attenuation is 55 decibels in steps of 1.5 decibels. An 8-channel control desk is shown in Fig. 8.

As previously mentioned, the studio control rooms have visible access to one or more studios. The second group of control rooms are without visible access to the studios and are designed primarily for connection to the output of the control rooms of the first-mentioned group, or to the speech studios, reproducing rooms, program sources outside of the house, the time signals, etc.

The control rooms of the first group are, therefore, used to monitor the various programs in the studios facing these control rooms, whereas the second group is used for the monitoring of all other program items. When the latter are used as main control rooms, the program from group one is passed through the control desk of group two without monitoring, this being done in the studio-control room.

5. *Amplifiers*

5.1 AMPLIFIER ROOM

The microphone and line amplifiers are mounted on standard-sized racks in the amplifier room, whereas the main amplifiers, loudspeaker amplifiers, etc., are located in their respective control rooms. In the amplifier rooms, there are two rows of 19-inch racks each consisting of 16 racks. These include amplifiers, receivers, oscillators, jack panels, remote panel for power plant, cable terminals, and similar equipment. The Broadcasting House is connected over a special 29-pair cable to a distributing center in the Telegraph Building, from where the lines to the transmitters are routed. For connection to various outside broadcasting points, a 100-pair cable has been laid to the nearest exchange of the Oslo Telephone Company. Both cables terminate in the amplifier room.

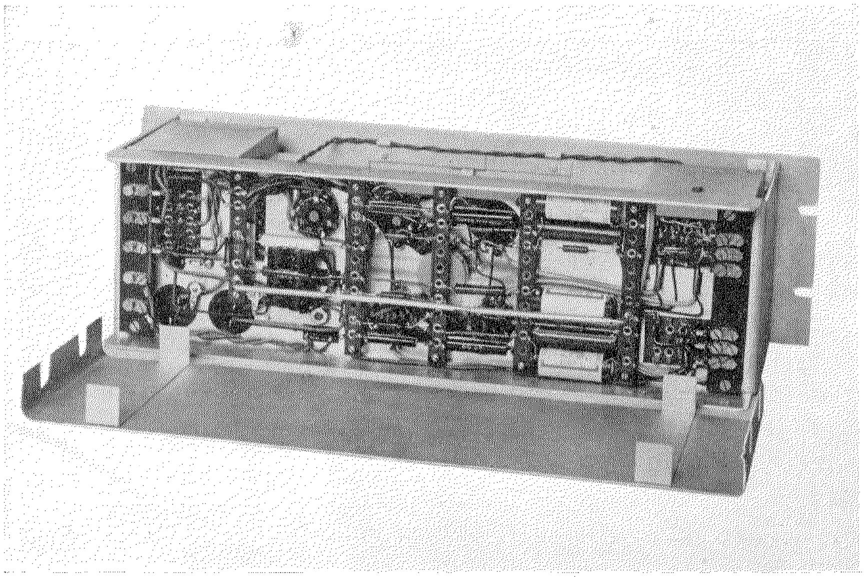


Fig. 9—Microphone amplifier, rear view.

5.2 MICROPHONE AMPLIFIER

The microphone amplifier, shown in Fig. 9, has a maximum gain of 60 decibels, adjustable in steps of 5 decibels down to 20 decibels by varying the amount of negative feedback. The response-frequency characteristic is within 1 decibel from 30 to 12,000 cycles per second. A maximum distortion of 0.2 percent occurs at 1.55 volts across a load resistance of 50 ohms. The noise level does not exceed 0.25 microvolt referred to a 50-ohm input. The microphone amplifier is also used

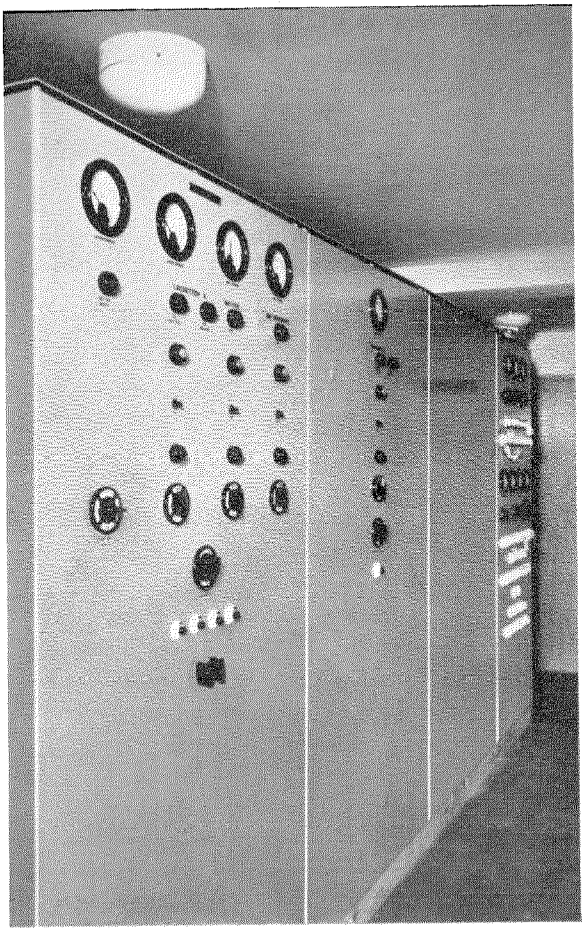


Fig. 10—Front panels of the rectifier units for energizing the automatic switching system.

as a main amplifier when limiting characteristics are not required.

5.3 MAIN LIMITING AMPLIFIER

The electrical characteristics of the main amplifiers are the same as for the microphone amplifiers. Limiting of output is provided for and this action starts at a minimum of 60 decibels below 1.55 volts (maximum output level). The output level will not increase more than 2 decibels for an increase in input level of 15 decibels.

5.4 LINE AMPLIFIER

The line amplifier has a maximum voltage gain of 8 decibels and is adjustable from −4 to +8 decibels in steps of 0.5 decibel. Response-frequency, distortion, and noise characteristics are the same as for the microphone amplifiers.

5.5 RECORDING AMPLIFIER

The recording amplifier is primarily a power amplifier mostly used for operating recording systems. It has a maximum output of 10 or 20 watts, which is produced by an input signal between 0.2 and 0.4 volt. Its response-frequency characteristic is flat within 1 decibel from 30 to 12,000 cycles. The noise does not exceed 5 millivolts across a 50-ohm output impedance. Distortion is less than 2 percent for rated output.

5.6 LOUDSPEAKER AMPLIFIER

The loudspeaker amplifiers are similar in characteristics to the recording amplifiers except that they provide for a maximum output of 20 watts. Noise will not exceed 10 millivolts across 50 ohms.

5.7 GENERAL

Push-pull circuits are used for all stages of all amplifiers. Negative feedback is employed for reducing distortion. Amplifiers are powered directly from the 220-volt, 50-cycle mains supply.

The amplifiers are of a design adopted by the Norwegian telegraph administration, who have also built the limiting, recording, and loudspeaker amplifiers in small quantities. A quantity of 130 microphone amplifiers has been delivered.

6. Power Plant

As all amplifiers are mains-operated through self-contained rectifiers, no central power plant is required for them. For the automatic switching system, direct current at 48 volts is provided by two selenium rectifier units, each with a regulated output of 35 amperes. Although normally used with a floating battery, each rectifier is equipped with automatic voltage regulation and can, accordingly, be used without the battery. The battery is required only in case of a mains-supply failure and supplies power only until an emergency diesel-engine generator is placed in operation. Fig. 10 shows the front panels of the rectifiers.

The speech-input equipment, including the switching system, control desks, microphone and main amplifiers, etc., was produced by Standard Telefon og Kabelfabrik A/S, in Oslo.

Portable Crystal-Controlled Frequency Standard

By R. I. B. COOPER

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DUPLICATE crystal-controlled frequency standards were studied under laboratory conditions and during the recent submarine gravity survey carried out in *H. M. S. Tudor*. A frequency accuracy within about 1 part in 10^7 was achieved during these trials.

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During August, 1946, a submarine gravity survey was carried out in *H. M. S. Tudor* with the object of obtaining information on the gravitational field over the continental shelf to the south and west of the British Isles. The measurements were made with a Vening-Meinesz pendulum apparatus. Gravity was determined at each station by comparing the period of a pendulum with that obtained at the base station where the value of gravity is known. To minimize perturbations due to the movement of the boat, the observations were made while the craft was submerged. Under these circumstances, an accuracy of gravitational measurement of the order of 1 part in 10^6 can be obtained, and it was necessary, therefore, to provide a means of timing the pendulum with at least this accuracy.

In the Vening-Meinesz apparatus actually employed, three pendulums were used, and their motions were recorded photographically by light beams reflected from mirrors attached to the pendulums near the knife-edges. Timing was done by interrupting the light beams at a rate controlled from the frequency standards.

The use of a submarine placed special requirements on the construction of the clocks. The first essential was that they should be small enough to pass through the 21-inch-diameter (53-centimetre) hatches in the boat. For general field use, they had to be constructed to withstand normal road and rail transport without special precautions. It was inconvenient to provide a thermostatically controlled oven for the crystal with its attendant bulky control equipment, so the crystals had to be capable of giving reliable operation over the temperature range from 10 to 30 degrees centigrade (50 to 86 degrees Fahrenheit), with only a simple correction for the ambient temperature as measured with an ordinary mercury thermometer.

The application of a precision quartz frequency standard to geophysical experiments in a submarine is not novel; similar equipment has been used by an American expedition.[1] Such a

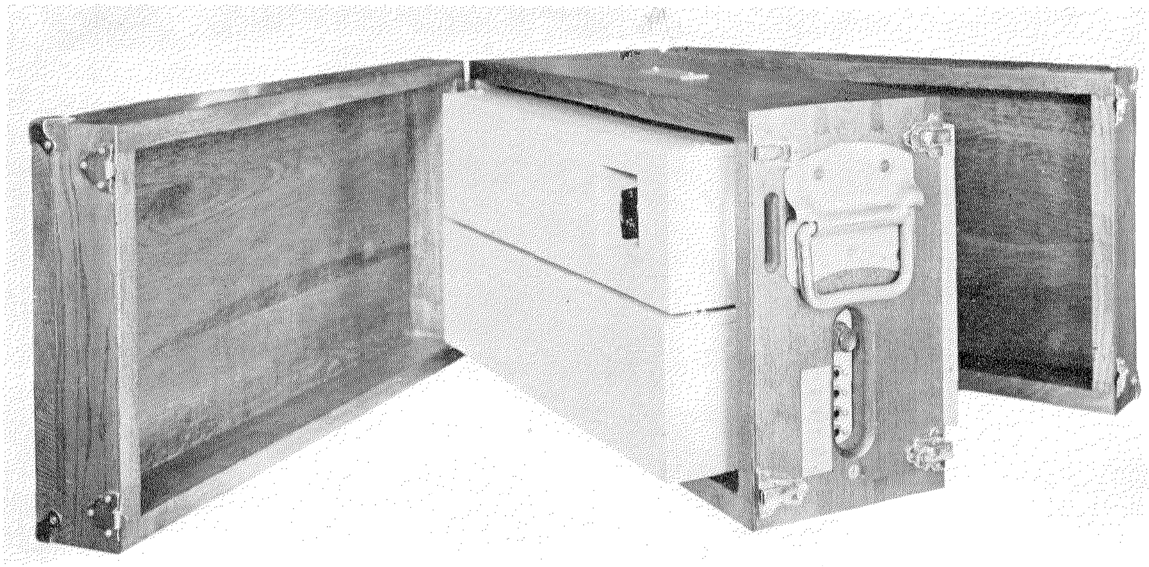

Fig. 1—Quartz-crystal-controlled frequency standard mounted in wooden box with both lids hinged.

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[1] Maurice Ewing, "Gravity Measurements on the *U. S. S. Barracuda*," *Transactions of the American Geophysical Union*, pp. 66–69; 1937.

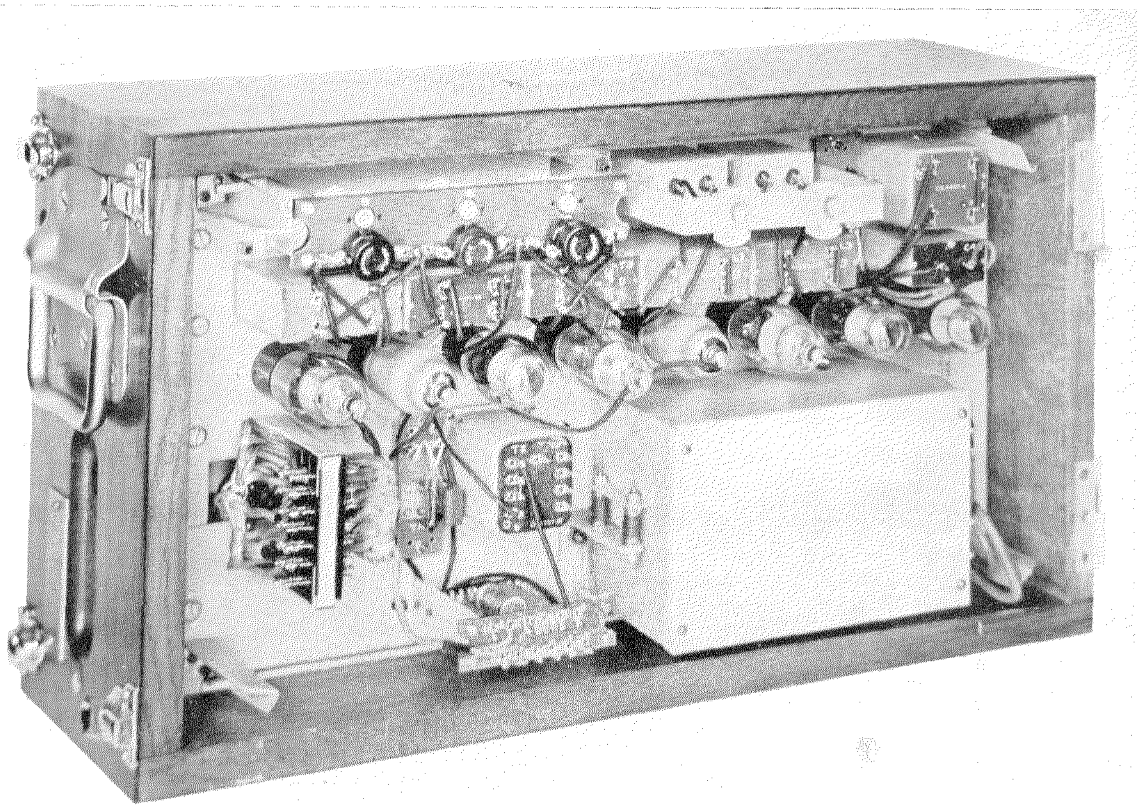

Fig. 2—Frequency-divider without protective metal covers. The crystal is mounted in the box in the lower right-hand corner.

standard was not, however, available in this country. It was known that Standard Telephones and Cables, Limited, had been making an extensive study during the war of high-precision oscillators, and it was, therefore, decided to approach them with a view to their undertaking the development and construction of the apparatus.

1. Construction

The complete frequency standard is mounted in a wooden box $22\frac{1}{4}$ by 15 by $12\frac{3}{4}$ inches, the total weight being approximately 100 pounds. The components are mounted on three steel panels, disposed on either side of the box, and access to their outer faces is obtained by opening hinged covers on both sides. On one side of the box are mounted the crystal-oscillator panel and power pack, and on the other side is the frequency-divider unit. Fig. 1 shows the box with the lid open and the apparatus panels protected by their metal covers. Fig. 2 shows the frequency-divider panel without its cover. The screened box in which the crystal unit is housed may be seen in the bottom right-hand corner.

2. Circuit Design

The oscillator operates at 100 kilocycles per second and is a modification of the feed-back bridge type developed by Meacham.[2] The oscillator was used in its original form by the previously mentioned American expedition. The crystal bridge is enclosed in a lagged chamber, which eliminates rapid fluctuations of temperature.

The crystal used is a *GT* cut, which provides the lowest temperature coefficient of frequency. The electrodes are formed by plating the face of the crystal, which is supported by connecting wires soldered to the electrodes at a nodal point

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[2] L. A. Meacham, "The Bridge-Stabilized Oscillator," *Proceedings of the I.R.E.*, v. 26, pp. 1278–1294; October, 1938: and *Bell System Technical Journal*, v. 17, pp. 574–591; October, 1938.

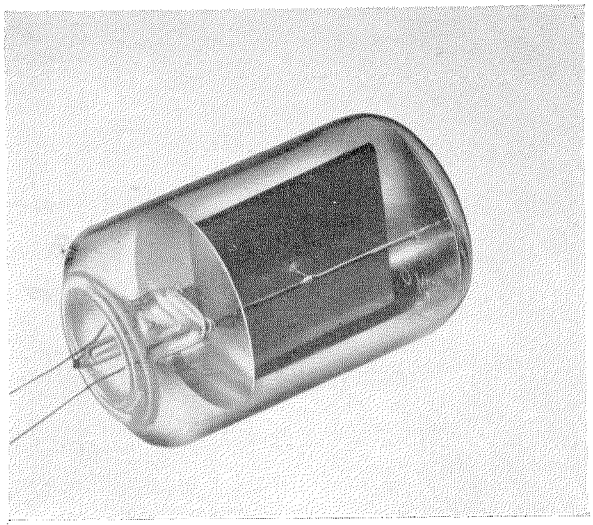

Fig. 3—Quartz crystal mounted in an evacuated bulb. The electrodes are plated on the faces of the piezoid or finished blank, and connecting and supporting leads are soldered to the electrodes at nodal points.

as may be seen in Fig. 3. The crystal is mounted in an evacuated glass bulb.

By lapping and aging, the temperature coefficient of a *GT*-cut crystal can be adjusted to a very small value, which remains constant over a considerable temperature range. This temperature range can be adjusted to occur, within limits, in a selected part of the temperature scale.

The initial frequency adjustment of the crystal is carried out to ±5 parts in 10^6, and the frequency of the oscillator can subsequently be brought within a continuously adjustable range of ±1 part in 10^6 by modification of the components of the crystal bridge circuit. This continuously adjustable range is under the control of the user.

The principal cause of frequency instability in a changing ambient temperature is the differential heating set up in the crystal. Hence, the crystal bridge is mounted inside a thermally lagged chamber, so that the rate of change of temperature to which the crystal is exposed is considerably reduced. A stability at a given temperature of the order of ±2 parts in 10^8 was measured in the laboratory.

The oscillator is stabilised against variations in supply voltages, and the output voltage is impressed on the frequency-divider unit. This unit provides high-impedance outputs at 10 and 1 kilocycles. The 100-kilocycle output itself is also available. These outputs are stabilised in voltage by a thermistor and a resistor connected across the output terminals. A thermistor is an element whose resistance diminishes with an increase of current; hence, variations in the current through it cause much smaller changes in the potential across it. A cathode-follower valve may be connected to any of these outputs whenever a low output impedance is required. The frequency divider operates on the harmonic regeneration principle. This is a continuously operating circuit that ensures that, short of actual failure of any component, the outputs will be controlled by the crystal. Each stage is arranged to divide by a factor of 10, and division by any other integer is very unlikely.

To increase the flexibility of the apparatus, it may be operated from either 200–250-volt 50–60-cycle alternating-current, 130- and 12-volt direct-current, or 12-volt direct-current power supplies.

In the laboratory and throughout the voyage, the sets were run from high- and low-tension accumulators continuously charged from dry rectifiers. The mains in a submarine are liable to occasional temporary failures.

3. *Performance*

Measurements were made, both in the laboratory and in the submarine throughout the voyage, of the instantaneous rates and the mean daily rates of the two standards. The instantaneous rate was obtained by comparing the 200-kilocycle harmonic with the carrier frequency of the British Broadcasting Corporation transmitter at Droitwich, which operates on 200 kilocycles. The beats produced between the two operated a counter, and the number of beats in 50 seconds was registered automatically. This figure is, therefore, the deviation from spot frequency expressed in parts in ten million. The corrections to be applied to the Droitwich frequency at the time of the observations were subsequently obtained, within an accuracy better than 1 part in 10^7, from the British Broadcasting Corporation from their own daily observations. Satisfactory checks of the frequency against Droitwich were obtained during the voyage up to a range of about 500 miles.

The 1-kilocycle output was amplified to a suitable power level and drove phonic motors, which produced time marks on the photographic paper used for recording the pendulums. These phonic motors also drove clock mechanisms. By comparing the readings of these clocks with the time signals transmitted at 1000 and 1800 hours from Rugby (GBR), it was possible to obtain the mean rate of each crystal over a protracted interval. The time signals were recorded photographically together with the clock indications, the reading accuracy being 1/61 of one second. This error is equivalent every 24 hours to 1.9 parts in 10^7. Corrections to the time signals were subsequently obtained from the Royal Observatory and applied to the readings where necessary.

The preliminary laboratory tests revealed that tilting, change of valves, and reduction in the supply voltage of 30 per cent caused no significant change of frequency. There was, however, a considerable frequency drift when the apparatus was first switched on, and this is shown in the curve of Fig. 4. This results from the operating temperature of the crystals being well above the ambient temperature.

During the voyage, over 80 observations of the instantaneous rates of each crystal were made and the ambient temperature at the time noted. The ambient temperature was taken to be that of the thermometer inside the pendulum apparatus, since it was considered that this apparatus and the frequency standards would have similar temperature characteristics. On plotting these results, it was found that the temperature coefficient of frequency of standard clock *2* was negligible, while that of clock *1* was -1.04 parts in 10^7 per degree centigrade.

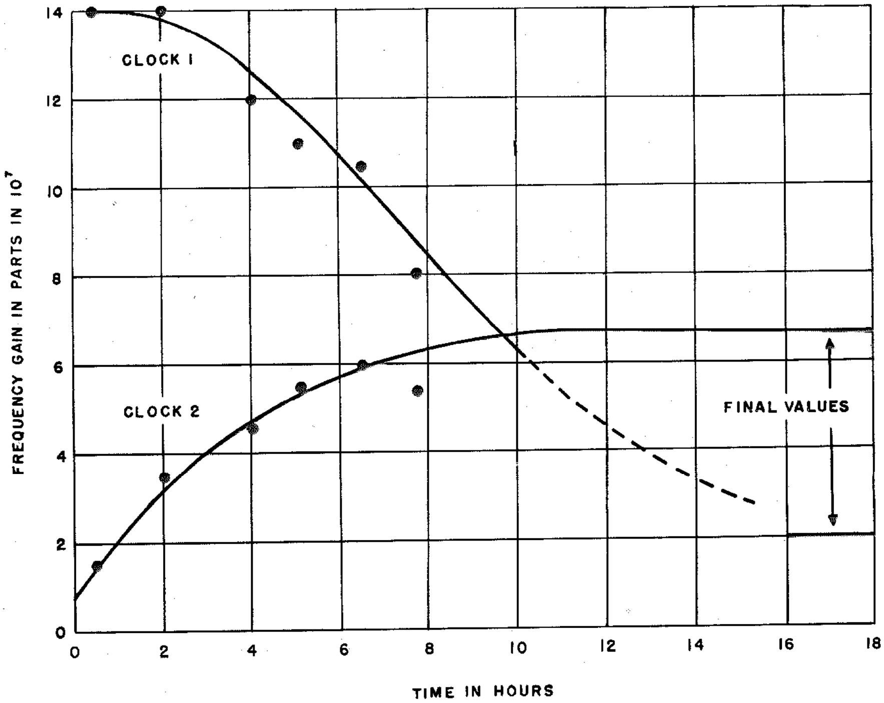

Fig. 4—Frequency drift during warming period immediately following switching on of equipment. Room temperature was 21.5 degrees centigrade (71 degrees Fahrenheit).

It should be pointed out, however, that although the ambient temperature appears to be a reliable guide to the behaviour of the crystals, the actual operating temperature of the latter was higher because of the heat dissipated inside the apparatus. Accordingly, these figures do not agree with the original coefficients measured in the laboratories.

The crystal boxes are at present being modified to permit insertion of a thermometer.

The variation in ambient temperature as measured above was found to be surprisingly small throughout the voyage, amounting to only 3 degrees centigrade (5.4 degrees Fahrenheit). To test whether the crystal frequencies were drifting, the results were reduced to a standard temperature of 20 degrees centigrade (68 degrees Fahrenheit), and weekly averages taken. These averages are shown in Table I, and it will be noted that there is no significant systematic change with time.

Less reliance is placed on the time-signal observations, and they are merely regarded as a useful check that no erratic behaviour was occurring. This is because the clocks were often stopped by accidental causes, such as blown fuses, and the period over which observations could be made was thereby reduced. The mean rates obtained over portions of the various weeks are, however, given in Table I for comparison. The agreement with the mean of the instantaneous rates determined from Droitwich is regarded as being satisfactory.

TABLE I

FREQUENCY DEVIATION IN PARTS IN TEN MILLION OF WEEKLY AVERAGES REDUCED TO A STANDARD TEMPERATURE OF 20 DEGREES CENTIGRADE (68 DEGREES FAHRENHEIT)

Date	Locality	Clock Number 1				Clock Number 2			
		Mean of Instantaneous Rates	Number of Observations	Rate by Rugby	Period in Hours	Mean of Instantaneous Rates	Number of Observations	Rate by Rugby	Period in Hours
23 June to 1 July	Cambridge	46.0±0.7	3	—	—	19.7±0.7	6	—	—
27 July to 2 Aug.	Portland and Channel Trials	45.9±0.3	10	—	—	20.6±0.15	11	—	—
5–9 Aug.	Atlantic Voyage	47.4±0.5	15	46.5	80	20.8±0.2	15	19.0	32
12–15 Aug.	Voyage to Spain	48.4±0.5	11	49.1*	48	20.5±0.2	11	20.9	48
18–24 Aug.	Voyage round Eire	47.8±0.5	12	46.6	80 32	20.7±0.2	12	21.0	80 32
25–31 Aug.	Oban to Portland	47.5±0.5	11	46.5	96	20.9±0.3	12	—	—
5 Sept.	Cambridge	48	2	—	—	20	2	—	—

* Failure of frequency demultiplier.

Throughout the voyage, there were only two major breakdowns in the timing system, one of which was directly attributable to the clocks themselves. This was due to a trimming capacitor becoming disconnected during rough weather in the Bay of Biscay on the 13th of August. The effect was that the output was no longer fully controlled by the crystal, and the 1-kilocycle output was changed in frequency by a few cycles. Since this was only a few cycles different from the nominal frequency, the clocks did not stop and the fault was not immediately apparent. This behaviour of the frequency dividers may well have been provoked by the development of a high internal resistance in the high-tension accumulators, which shortly afterwards were the cause of a second breakdown.

The internal resistance of the high-tension batteries became so high as a result of corrosion that about 10 volts (root-mean-square) of ripple from the charging circuit appeared in the supply. The effect was to change the 1-kilocycle output slightly in frequency. This may have been due to modulation of the oscillator frequency, causing the divider to operate on one of the modulation products instead of on the oscillator frequency itself. It is probable that the gradual development of this fault was responsible for the high mean rate recorded for clock *1* between the 12th and 15th of August.

It is clear that, despite the variations with temperature mentioned above, the sets fall well within their original specification, which was for an accuracy of 1 part in 2,000,000. They also fulfill excellently all the other requirements.

4. Acknowledgment

Thanks are due to Messrs. Standard Telephones and Cables, Limited, for the skill and care given to the development of these instruments and for providing the technical information on which this account of their operation is based.

The author is also indebted to the Department of Scientific and Industrial Research for a grant that made possible his participation in these researches.

Throughout this work, I have had the advice and encouragement of Mr. B. C. Browne, who was in charge of the arrangements for the expedition.

Remote Control*

By C. GORDON WHITE

Standard Telephones and Cables, Limited, London, England

THE rapid extension in the application of supervisory remote-control schemes at home has been paralleled by its ever increasing use in many parts of the world. Its main sphere of development, as would be expected, has been in the control and supervision of electrical networks, not only those serving densely populated metropolitan areas, but also other schemes, such as electrified railway systems and the transmission and distribution substations of large electrical networks.

With the extending use of remote control has grown also a greater demand both for the facilities to be provided and the degree of flexibility to be given, without appreciable increase in cost. To meet such needs, it has been necessary to employ to the full techniques peculiar to the telephone communications field rather than those of the switchgear world, and it is not surprising to find that the principles of signal transmission and automatic switching peculiar to telegraphy and telephony, as also the apparatus specially developed for these purposes, are being increasingly used to satisfy present-day control needs.

It can readily be seen that there is a marked similarity in the two operations, namely, the selective ringing of a particular subscriber's telephone bell and the remote operation of a particular switchgear unit in an electrical network. The means of carrying out both these operations have much in common, more especially when similar apparatus is employed, but in the

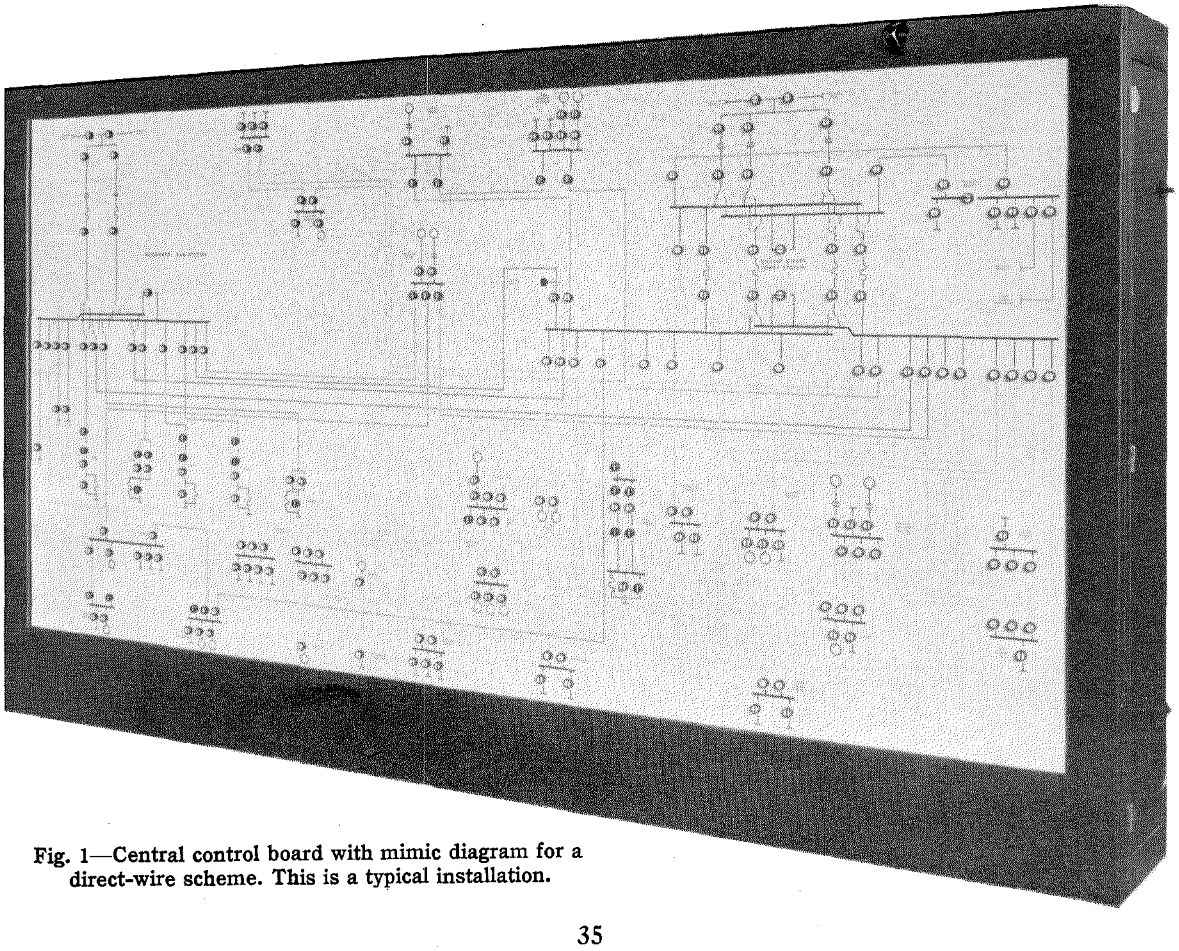

Fig. 1—Central control board with mimic diagram for a direct-wire scheme. This is a typical installation.

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* Reprinted from *British Engineering Export Journal;* March and April, 1946.

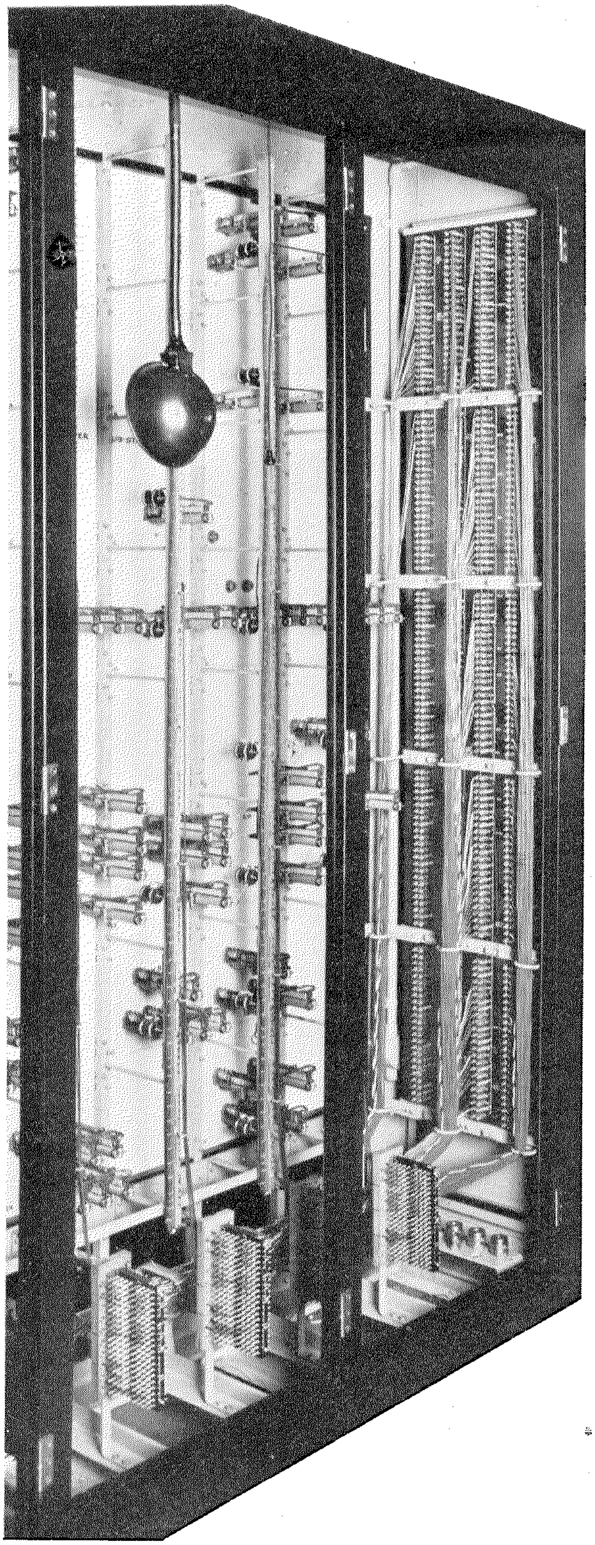


Fig. 2—Rear view with the doors removed of the central control board shown in Fig. 1.

approach to the two problems there are marked differences. In both, economic considerations for the most part determine the quality of the complete equipment and the degree of accuracy in the final desired result. In telephone switching, an occasional lost call or wrong number can, on this account, be at least tolerated, but not so in the case of the control of switchgear units where it is imperative that a wrong operation just must not happen, even if the alternative to this is that nothing whatever happens. The operating conditions normally experienced help greatly towards this end, since the variables obtaining in telephone switching, such as line resistance, impulsing speed, and rates can be closely controlled, but even so, special precautions are taken in the design of supervisory equipment to safeguard signals against possibility of mutilation and to ensure correct operation.

1. Substations Within One-Mile Radius

1.1 SIMPLE DIRECT-WIRE CONTROL SCHEME

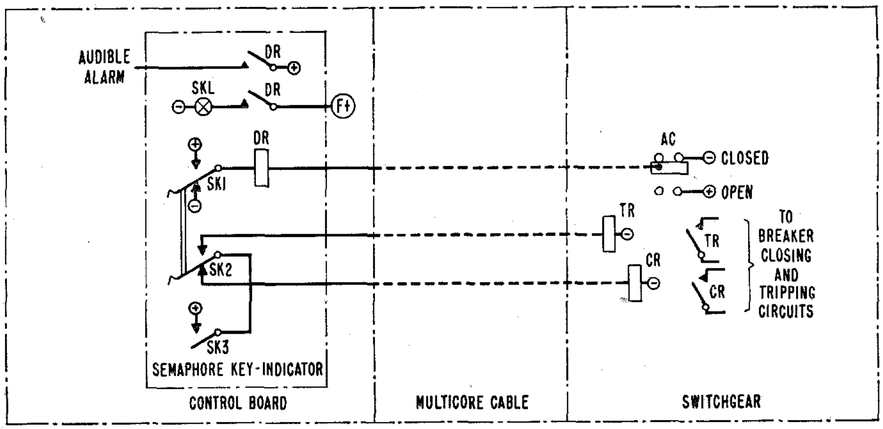
For the supervisory control and indication of electrical substations within a radius of approximately one mile, it is often possible to lay down multi-core cable and operate on a direct-wire basis. Then with a common-battery supply, the overall scheme can show an appreciable saving in first cost, whilst providing the definite advantages accruing from initial simplicity. Heretofore such schemes have usually operated at 110 or even 220 volts over power-type multi-core cable, but with the alternative using light-current apparatus, working at, say, 24 or 50 direct volts, it is possible to employ multi-core telephone-type cable incorporating copper conductors of 10 or 20 pounds (4.5 or 9 kilograms) per mile, being equivalent respectively to 23 and 20 steel wire gauge (0.61 and 0.91 millimetres), and show a marked saving in cost without any loss in reliability or robustness. The type of cable must depend upon the local conditions and its proximity to other electrical circuits, but in general it should be lead-covered, with steel tape or wire armouring. Such schemes sometimes employ ingenious circuit arrangements relying upon marginal currents to reduce the total number of cores, but the use of telephone-type cable permits of the relatively generous use of conductors with consequent simplicity and sturdiness in design. A typical installation of equipment operating on these principles is illustrated in Figs. 1 and 2, whilst Fig. 3 shows an outline of the simple schematic arrangements on which the system is based.

This particular project concerns a group of electrical substations serving a large steel works and covering a wide area, which are interconnected with the central control point by multicore telephone-type cables. Each circuit-breaker or isolator is represented on the mimic-diagram control board by a combined key-indicator of the semaphore type (see Figs. 4 and 5). This unit, which is one-hole fixing, incorporates a miniature lamp (behind a translucent insert), which lights up or flickers when the indicator and its corresponding switch at the remote end are "out of agreement"; and in this way visual indication (accompanied by an audible alarm) is given when the distant unit changes its position, e.g., a breaker trips on fault. The semaphore serving as a mechanical position indicator need not normally be illuminated, an arrangement which gives a greatly increased life to the miniature lamps, and has proved a decided advance upon similar schemes where continuously illuminated lamps are employed.

It will be noted that the diagram board itself is sectionalized to facilitate subsequent network modifications and extensions. Some idea of the construction of the cubicle and the internal wiring layout may be gleaned from Fig. 2. It will be noted that the large-type cable-forms usually associated with such equipments are obviated and that loose "jumper" wiring is used to connect the indicator circuits with the terminals in the base of the cubicle. An end view of the cubicle is shown in Fig. 4, where behind a tight-fitting hinged door are located the isolating links associated with each incoming conductor and provided to facilitate maintenance.

Reverting to the diagram of Fig. 3, the semaphore key-indicator *SK* is shown in the "closed" position corresponding to auxiliary contacts *AC* on the remote circuit breaker. When the breaker trips, contacts *AC* change over and positive battery is fed via *AC*, the pilot wire, relay winding *DR*, and contacts *SK1* to negative, thus energizing relay *DR*. This relay flickers the semaphore lamp *SKL*, and also rings the audible alarm. When the signal is "accepted" by turning the semaphore key through 90 degrees to the "open" position, the winding of *DR* is short-circuited and normal conditions are restored.

Where control facilities are included, the interposing relays at the distant substation are connected as shown and the semaphore indicator is fitted with spring-loaded "overturn" positions.



NOTES -

SK1 AND SK2 OCCUPY EITHER OF TWO POSITIONS ACCORDING AS KEY INDICATOR IS IN THE "OPEN" OR "CLOSED" POSITION.
SK3 CONTACTS CLOSE WHEN THE KEY IS OVERTURNED FROM EITHER THE "OPEN" OR "CLOSED" POSITION.

CR = "CLOSE" INTERPOSING RELAY.
TR = "TRIP" INTERPOSING RELAY.
SK = SEMAPHORE KEY INDICATOR.

DR = "OUT OF AGREEMENT" RELAY.
SKL = LAMP IN SEMAPHORE KEY.
AC = BREAKER AUXILIARY CONTACTS.

(F+) = INTERRUPTED "FLICKER" SUPPLY

Fig. 3—Direct-wire scheme for a group of substations serving a large steel works.

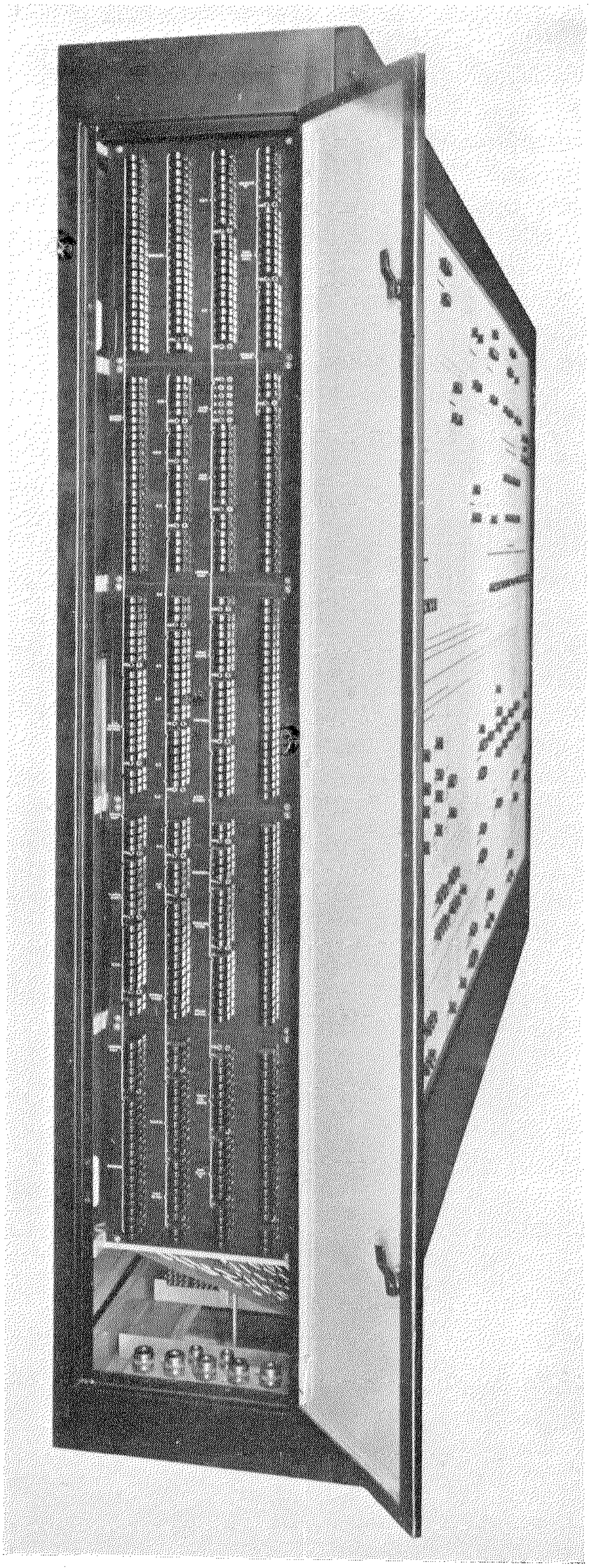


Fig. 4—End view of the central control board (see Figs. 1 and 2) showing the conductor isolators.

2. Pilot Wires and Separate Frequencies

2.1 PILOT-LINE NETWORKS

Where the use of direct-wire schemes are prohibited by distance, cost, or the difficulties of excavation for cable, resort must be had to other methods employing the minimum number of pilot wires or, alternatively, using separate frequencies superimposed on circuits already existing for other purposes.

Dealing first of all with the former case, the final arrangement depends very much on local conditions, the layout of the stations involved, and the range of facilities to be satisfied. Three types of station grouping about a central point are indicated in Fig. 6. That at (*A*) represents the connection of a number of stations to the central point on a radial network, i.e., one pilot circuit separately to each remote station. At (*B*) is shown a number of distant substations grouped radially about the intermediate, parent, or tandem station *T*, which in turn is connected to the control point *CS* by an independent pilot line. This method is economical in conductors between the tandem and central stations, particularly on very large schemes involving hundreds of stations, but may result in slight delay in transmission of signals to the central point if more than one substation in any one group wishes to signal at the same time.

Yet another layout is that of (*C*). This is known as the omnibus or party-line arrangement in which one through circuit feeds a number of substations connected thereto in parallel. This represents an appreciable economy in pilots where the geographical layout and the degree of traffic permits of its use and without any deterioration in the quality of the service to be obtained from the equipment. Obviously, it is possible to plan pilot-network layouts combining the arrangements outlined above, and this is usually the case in large schemes of centralized control.

The pilot lines referred to above usually consist of one pair of conductors for signal transmission in both directions for control and indication. Other facilities such as telemetering can also operate over the same wires or with greater simplicity can be accommodated on additional conductors on the same route.

In the older, densely populated countries, the provision of separate pilot circuits for central-control schemes presents little difficulty, especially where such circuits can be conveniently rented from the national postal and telegraph administration. Even here, however, the ever-widening ranges of facilities demanded necessitate either more and more metallic conductors or further inroads on the existing conductors by using superimposed circuits.

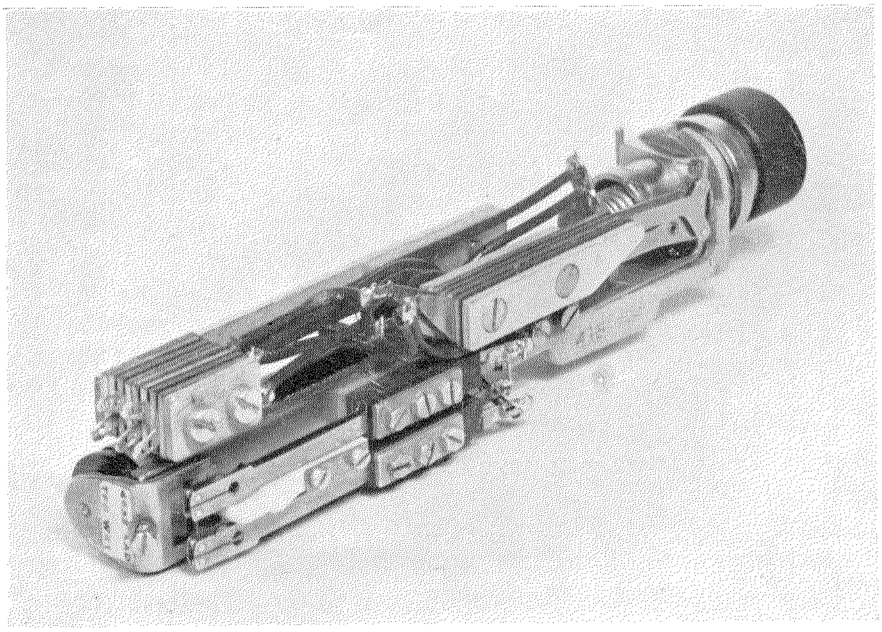
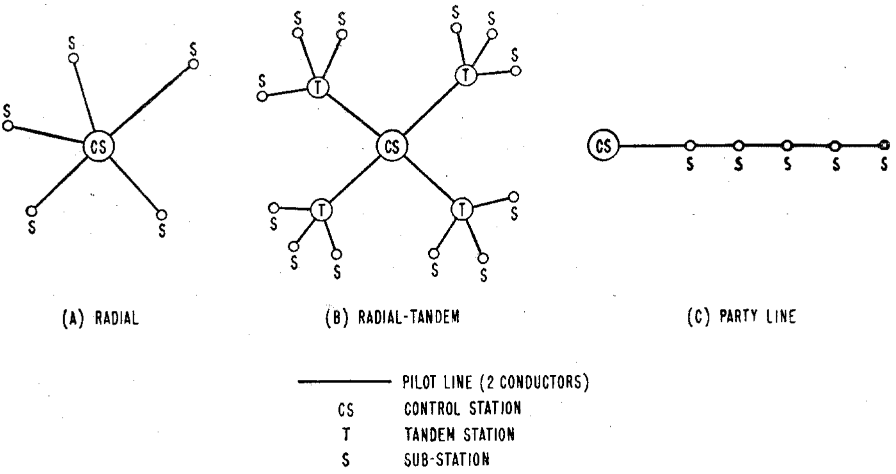


Fig. 5—Semaphore key-indicator with associated relay. It is a one-hole fixing for control boards.



(A) RADIAL (B) RADIAL-TANDEM (C) PARTY LINE

Fig. 6—Typical pilot-line arrangements for three station groupings about a central point.

3. *Telegraph Carrier Transmission*

Developments in the fields of telegraph carrier transmission suggest a ready solution to the problem, since it is possible to provide hundreds of independent circuits simultaneously on a suitable metallic-line circuit. In many instances, however, such line circuits just do not exist physically, and then provision for these facilities alone may not be warranted, but where a high-tension line connects the controlling and contolled stations, resort may be had to carrier circuits superimposed on the high-tension line itself. The main expense in such schemes is in the provision of the terminal equipment, more especially the apparatus which separates the high-tension power circuits from the signalling equipment, but for long distance across country in areas either undeveloped or unsuitable for separate pilot circuits, these prime costs may well prove very small compared with the advantages to be gained by the use of centralized control.

3.1 Typical Signalling Scheme

A general outline of a scheme for control and indication of an electrical substation from a central point can be gathered from the block schematic of Fig. 7. For simplicity, let us consider a single station served by a line of two pairs (four conductors). Both locations are equipped with apparatus for sending and receiving signals consisting of trains of impulses coded to represent distinctive numbers, much the same as multi-digit numbers when generated on an automatic telephone dial. At the control station, most of the signals are initiated as a result of a definite manual operation by the control engineer, whereas most of those issuing from the substation originate automatically, due to involuntary movement by individual mechanisms.

For instance, should a breaker trip, say, on fault, its auxiliary contacts via its "link" circuit stimulate the substation code sender which, seizing the line, transmits a signal (lasting approximately two seconds), which is routed at the control station to the code receiver. This apparatus sorts out and checks the message and directs it in turn to the appropriate part of the equipment, in this case the control-board indicator corresponding to the switchgear unit concerned. In the same way, signals from the control station are generated by the code sender there, which seizes the line, and the resultant impulse train passes to the code receiver at the substation where it is "de-coded" and directed to the apparatus concerned.

It is here that the safeguards referred to at the beginning of this article are of importance. One method to ensure that only correct operations are effected is by the use of signal codes specially compiled. Each code may consist, say, of three numbers or digits, *A*, *B*, and *C*; but all codes have one thing in common, that *A* plus *B* plus *C* is a constant value. On receipt of any code at the distant end, it is checked for the correct number of digits and also the correct total, and only when these conditions are satisfied can this code be effected.

Another method adopted is that of "preselection checkback." Here, before a control operation can be effected, a preliminary or preselection code is sent down the line to "pick out" the unit which it is desired to control. When this selection has been done, a "check-back" code peculiar to the selection made is sent to the control station and, if it "agrees" with the originating unit, a special "go ahead" signal is given to the controller. The latter then completes his final operation whereupon a further signal is sent down the line to be followed by the appropriate position-change indication back to the control station, confirming that the desired operation has been completed. The pre-selection condition previously set up is then cleared down, leaving the line free for other signals.

4. *Advantages of Signal Codes*

The first-mentioned method has decided advantages, not only in simplicity and speed of operation, but also because it "occupies" the pilot line only during actual signal transmission,

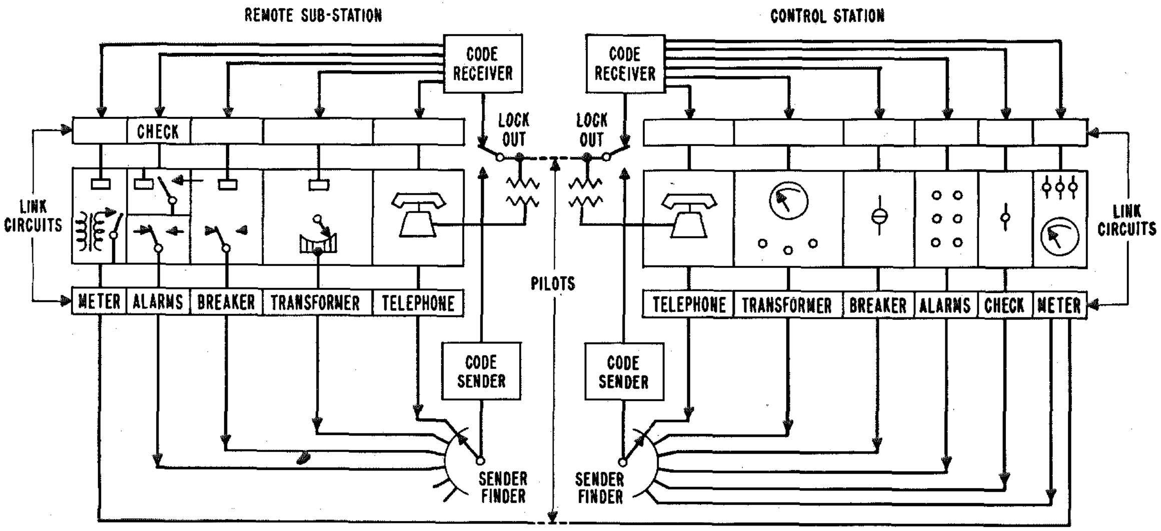


Fig. 7—Block schematic for control and indication of an electrical substation from a central point.

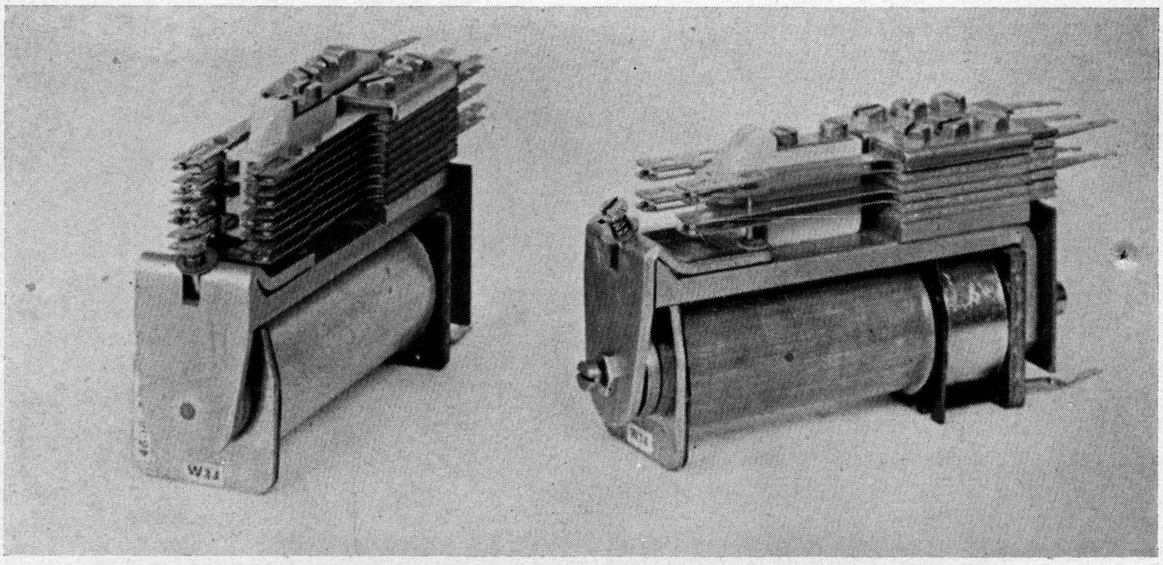


Fig. 8—Telephone relay, generally mounted in groups according to circuit requirements.

whereas with "preselection check-back" the line is held from the initial preselection operation until final clear down, thus preventing its use for the transmission of other signals that may require urgent attention.

An alternative to the use of coded signals, and one that has been widely used, employs rotary distributors at each location. These are rotated simultaneously, checks being provided to ensure that they keep in step or are otherwise correctly synchronized in each possible position. Such schemes require, in general, a greater number of pilot-wire conductors and cannot supply the same high degree of flexibility as is inherent in schemes employing coded signals.

4.1 Apparatus

Systems based on telephone practice employ in the main relays, selectors, and suchlike apparatus, as have been standardized for telephone use, and these in vast numbers have proved their reliability in the very exacting field of constant daily operation on telephone systems in all parts of the world. The principal items employed comprise relays (Fig. 8) and uniselectors (Fig. 9), which are generally mounted in groups according to circuit requirements on multi-point jack-in panels, similar to that shown in Fig. 10. Where there is a number of similar circuits on one set of equipment or a number of substation installations generally similar to one another, this type of mounting gives the maximum flexibility and greatly facilitates normal maintenance work. A few spare mountings of each kind can be held in readiness for substitution as need arises, thus enabling a faulty unit to be returned to a central depot for repair under the best possible conditions.

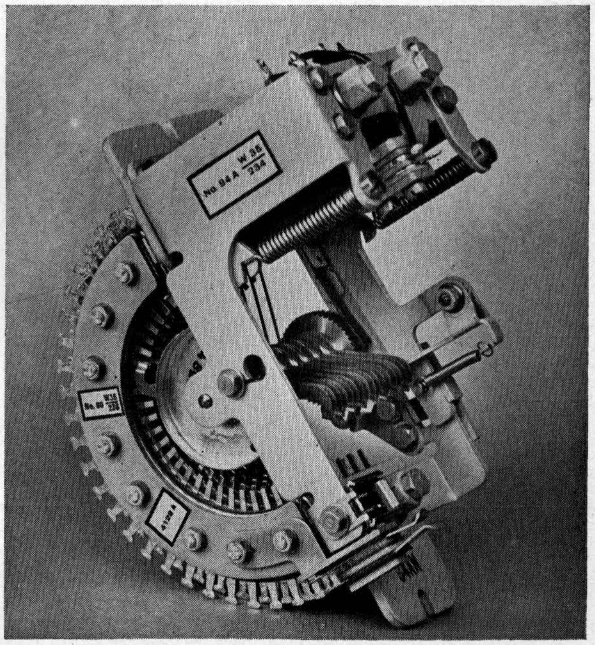


Fig. 9—Uniselectors are among the principal items of apparatus standardized for telephone practice.

5. *Appearance of Control-Room Equipment*

For the equipment located in control rooms and mounted on control boards and desk, it is particularly desirable to improve on the normal telephone switching product so that these equipments should present a really good appearance, having a high-grade finish and carried out in colour tones pleasing also to the eye. There is increasing appreciation of the fact that the working surroundings play a very important part in ensuring contented operators and efficient operation, and the quality of workmanship of control panels and desks and their artistic design can contribute greatly to the desired end. It is not, therefore, sufficient to specify loosely that the equipment should mount keys and lamps, but some idea of the quality of workmanship should also be laid down so as to encourage those manufacturers who pay particular attention to these features and are willing to initiate special designs to give the best possible appearance to their products consistent with the conditions in which they are to be used.

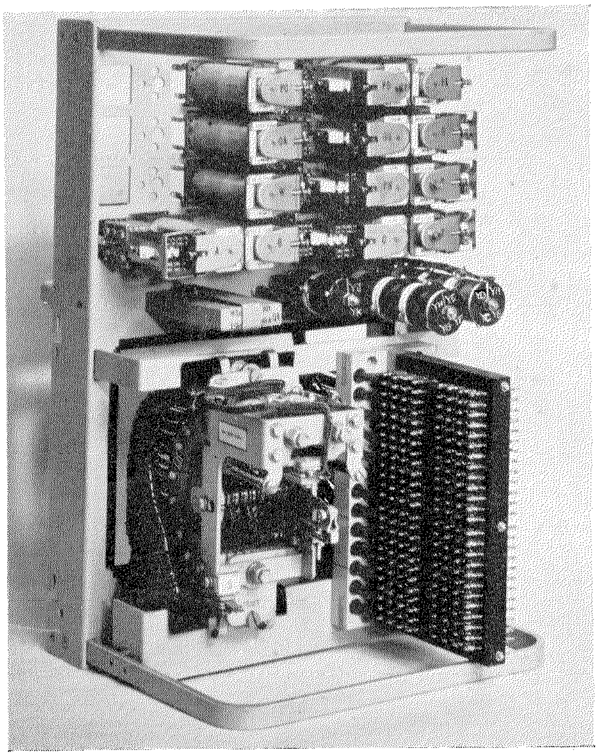


Fig. 10—Jack-in type apparatus panel with cover removed. This type of mounting gives maximum flexibility and facilitates maintenance.

6. *Centralized Supervisory Control Schemes*

Table I lists a selection only from the many supervisory remote-control and indication equipments designed and manufactured by Standard Telephones and Cables, Limited. Each installation is served from a central control room. The facilities totals, though approximate, only serve to indicate the size and range in each case.

TABLE I

Supplied to	Location	Details
Central Electricity Board	Mid-East England Area (Leeds)	31 remote stations; indication of 68 switches and 30 transformers; 55 remote meter readings; 9 total load readings; 9 routine instruction panels; automatic and manual switching, priority and emergency telephone facilities.
Wellington City Council	Wellington, New Zealand	24 stations; 190 switches and transformers; 16 remote meter readings; mass firing group control; miscellaneous alarms telephone; remote routine testing.
Great Indian Peninsular Railway	Bombay Kirkee-Igatpuri (Kalyan)	14 remote substations and track cabins; control and indication of 220 switches; miscellaneous alarms.
London Passenger Transport Board	Underground: Northern and Western extensions (East Finchley and Old Oak Common)	15 rectifier substations; 450 switches controlled and indicated; 53 meter readings (voice-frequency carrier system); 60 miscellaneous alarms.
Manchester City Corporation	Manchester	Common-diagram system; centralized control of city networks (33 kilovolts, 11 kilovolts and 6.6 kilovolts), 95 substations, 500 switches and transformers controlled and indicated. Planned for ultimate extension to 400 city distribution stations.
London, Midland & Scottish Railway	London Euston-Watford Line (Stonebridge Park)	Common-diagram system: 17 rectifier substations and associated power station, 200 breakers, 240 alarms, telephones.

Control Station Apparatus*

By C. GORDON WHITE

Standard Telephones and Cables, Limited, London, England

A TYPICAL example of a modern control station equipment is illustrated in Figs. 1 and 2, which show a central control desk supplied for a large city network in the Dominions. In this case, the main part of the city electrical network is shown on a large curved wall diagram approximately 30 feet (9.1 metres) long and 11 feet (3.3 metres) high to the rear of the central control desk. The wall diagram is made up of 234 panels each 10×10 inches (25×25 centimetres), which mount automatic semaphore indicators incorporating miniature lamps, manual indicators, alarm lamps, etc. A close-up view of part of this diagram is shown in Fig. 3. The desk itself carries a restricted mimic diagram incorporating control key-indicators of the semaphore type solely for the switchgear units subject to remote control. Towards the back of the desk diagram is fitted an alarm-locating indicator to facilitate the detection of the source of any alarm by the operator. When a breaker, say, trips, the audible alarm sounds, but at the same time this indicator shows the name of the station of origin and the type of signal that has been received. The operator is thereby able to direct his attention immediately to that part of the wall diagram with which the change is associated, and thus complete his duties most rapidly and efficiently.

These indications are signalled on miniature lamp facias incorporating photographic plates carrying the various legends. The letters are normally invisible, but show up clearly in white or coloured lights when they are illuminated.

Other examples of typical control boards are seen in Figs. 4 and 5. The former, comprising a desk in waxed oak mounting a mimic diagram for seven distant high-tension stations, concerns the control of a system operating over a mountainous pilot route. Fig. 5 consists of a single isolated control panel for a pumping station and includes meter readings of electrical load and water level.

It will be noted that in these examples care is taken in the mounting of keys, meters, etc., to avoid so far as possible the use of screw heads that may be visible from the front of the panels. In Fig. 5, the meters, be it noted, are mounted at the back of the panel behind a circular bevelled cut-out, which serves to give a pleasing and smooth appearance to this part of the board.

It is well-nigh essential on such schemes to pay particular attention to the way in which the apparatus is to be housed. The conditions in substations vary very greatly, both as regards temperature, humidity, and general cleanliness, and in many cases represent much more severe conditions than those in which telephone equipment is normally housed. It is difficult to improve upon the use of pressed steel cubicles, arranged for floor or wall mounting, where special precautions are taken to exclude dust. Typical examples of such equipments are shown in Figs. 6,

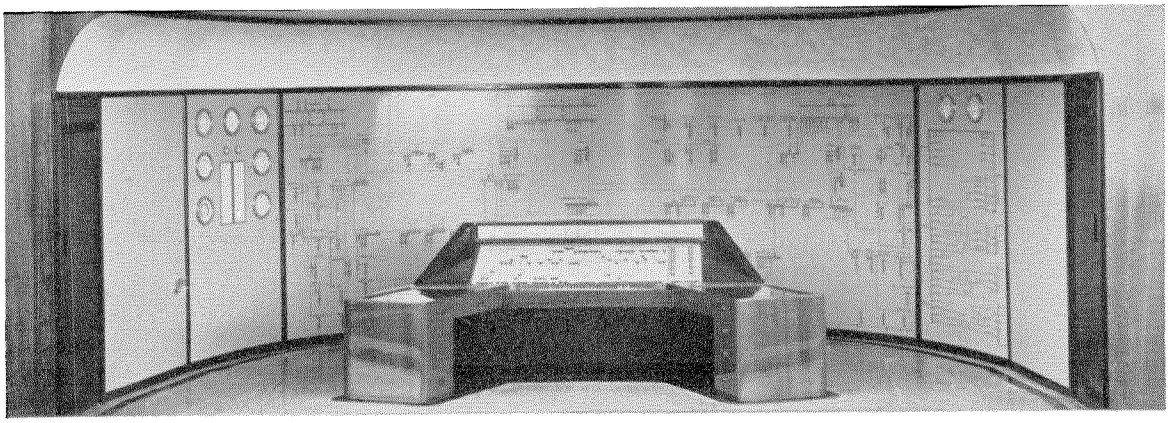


Fig. 1—Control desk and curved wall diagram for a centralized network of a large city in the Dominions.

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* Reprinted from *British Engineering Export Journal;* March and April, 1946.

7, and 8. The first mentioned, a floor-mounting unit, is fitted with hinged doors, which fit snugly on to rubber insertions to give a dust-tight enclosure. Fig. 7 shows a small wall-mounting unit, the method of hinging to give access to the rear wiring being shown in Fig. 8. Towards the top or base of each cubicle are fitted loose plates, which can be drilled on site as necessary to take the cable terminations, which assists to give a sealed enclosure. Where the conditions warrant it, local heater circuits can readily be fitted, and in humid conditions readily detachable air-drying units can be incorporated.

In the substation cubicles are mounted the terminal boards connecting the small internal apparatus wiring to the stouter conductors incoming from the switchgear units. One such type is illustrated in Fig. 9 and is similar to those already to be noted in Figs. 6, 7, and 8. This terminal block is a moulded type of robust design and can incorporate lamps where necessary, or, for instance, a fuse as is shown in this illustration. Each outgoing connection is terminated on an *OBA* stud, being served by an isolating *U*-link with insulated handle. This ready means of isolating these circuits is invaluable during maintenance work.

For equipment destined for tropical conditions, special precautions are taken with regard to apparatus finish, design of wiring, and type of enclosure. Manufacturers when designing such equipments are ever alive to the special needs in this respect, and further advances in tropical finishes can be confidently anticipated in the light of the varied and widespread experiences of the Services during the war.

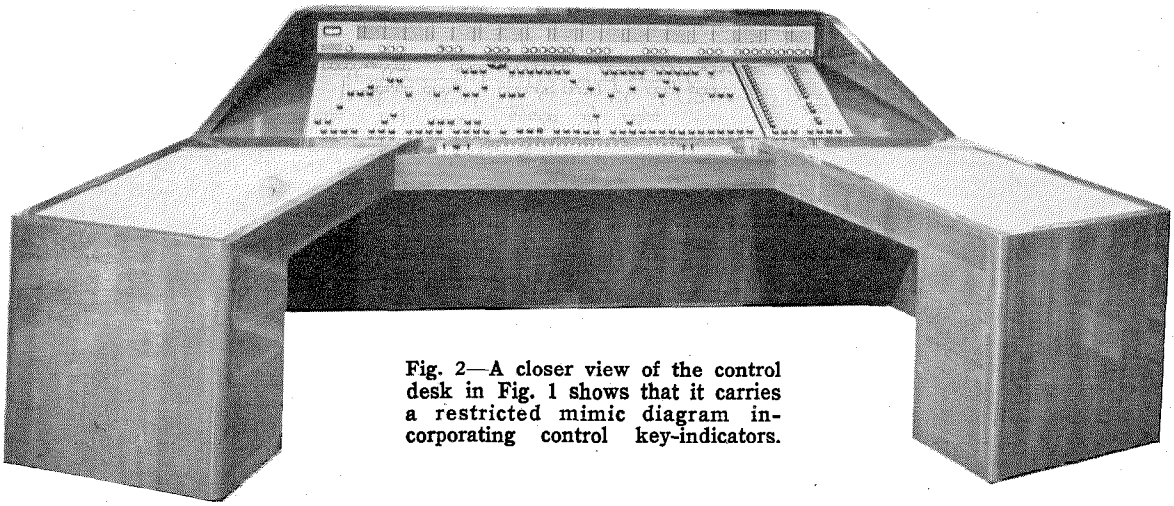

Fig. 2—A closer view of the control desk in Fig. 1 shows that it carries a restricted mimic diagram incorporating control key-indicators.

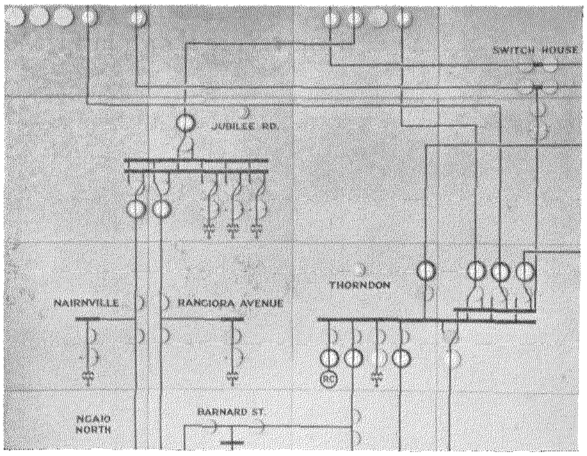

Fig. 3—Wall diagram detail from Fig. 1. The complete diagram, which is curved, is 30 feet (9.1 metres) long and 11 feet (3.3 metres) high, and is made up of 234 panels 10 inches (25 centimetres) square.

Among the facilities stipulated in connection with planning of central control schemes are those of giving various meter readings at the central point. A number of simple and ingenious schemes have been devised to meet such requirements, a few only of which can be touched on here.

1. Simple Loading Indications

For simple loading indications, where the distances are short and a high degree of accuracy is not imperative, direct metering offers a ready solution. Here (see Fig. 10), for readings of

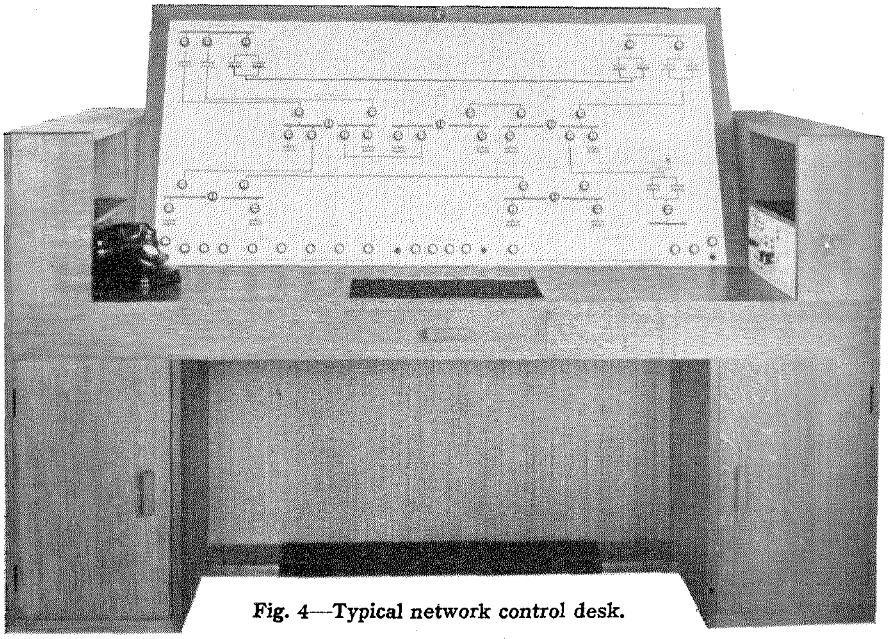
Fig. 4—Typical network control desk.

alternating current or voltage, connection is made over the pilots between the current or voltage transformer at the substation and an indicating milliammeter at the control station, suitably calibrated to make allowance for the pilot resistance. If possible, a separate pilot pair for this purpose permits of such an indication circuit once selected being maintained as long as may be required, without affecting in any way the retention of the main signalling circuit for all control and indication purposes.

If spot readings of this type are adequate, both the signalling and metering facilities can be provided simply over the same pair. For this, "slow-acting" relays are included at each end of the line, and when such a "timed" meter selection is initiated the line is switched temporarily by these relays from the supervisory equipment to the selected metering circuit. The period for which the reading is available is adjustable, and may be of the order of 15–20 seconds. Meanwhile, supervisory signals, which may be initiated, are automatically stored and transmitted later as soon as the meter switching relays return the line again to its normal use.

Where greater accuracy is required, or longer lines, not necessarily metallic throughout, are experienced, resort must be had to other means of obtaining quantitative readings. One such method for transmitting various types of meter readings over a telephone channel is available in the Variable-Frequency Continuous (VFC) metering system. One form of this is illustrated schematically in Fig. 11, a typical transmitting meter being shown in Fig. 12. For this scheme, the transmitter—a meter of otherwise normal design and type—is fitted with a small variable capacitor connected in the tuning circuit of a variable-frequency oscillator so that the

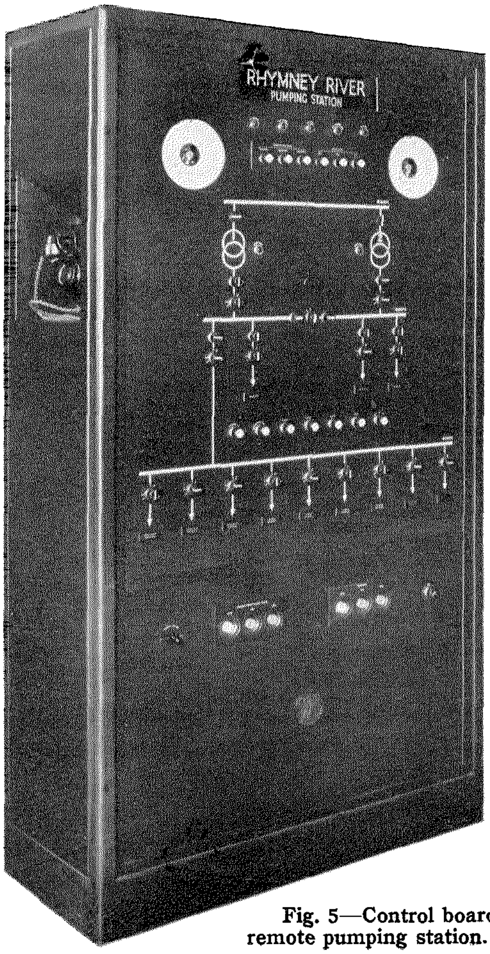

Fig. 5—Control board for remote pumping station.

frequency output is a function of the meter reading. The variable voice frequency thus generated, on receipt at the distant end of the line, is amplified, cleaned of harmonics, limited in peak voltage, and is then passed through a discriminating network, the output of which in direct current is inversely proportional to the frequency input. The direct-current indicating meter connected in this output circuit is calibrated to correspond with the distant transmitting meter, and follows faithfully and without any appreciable time lag the fluctuations of the corresponding transmitting meter. This system is of course by no means limited to transmission of electrical quantities can readily be used for a variety of readings, such as air pressure and the like.

The frequency band corresponding to one meter circuit is only 50 cycles, so that by the use of filters several such circuits can be accommodated simultaneously on a channel otherwise suitable for normal speech transmission.

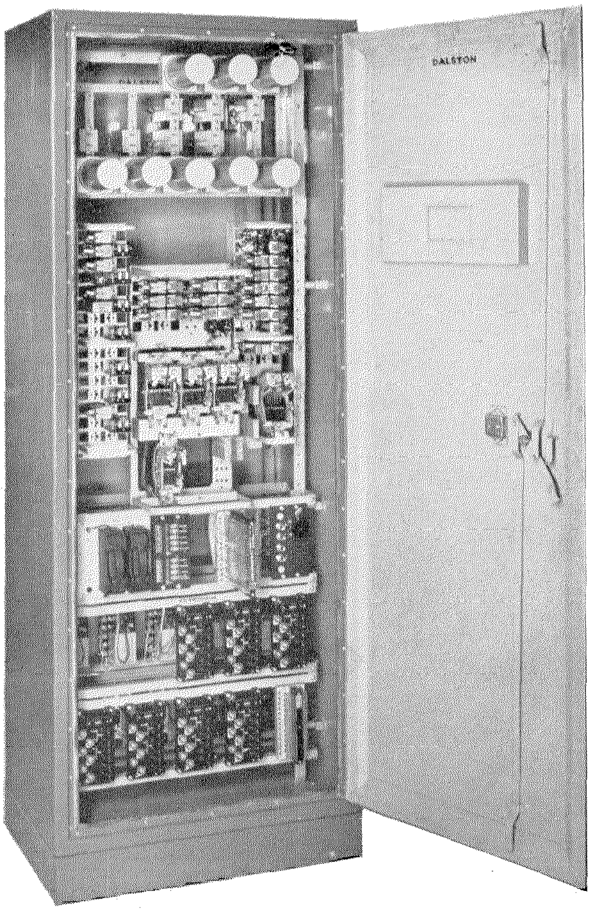

Fig. 6—Substation apparatus, cubical for floor mounting.

An alternative to this is the "photo-telemetering" system. In this a voice frequency, generated by the transmitting meter incorporating a small motor and photo-electric cell, is interrupted at intervals to give a series of impulses, the ratio of "make" to "break" of which is a measure of the reading to be transmitted. This is received ultimately on a milliammeter suitably calibrated to agree with the scaling on the associated distant transmitting meter.

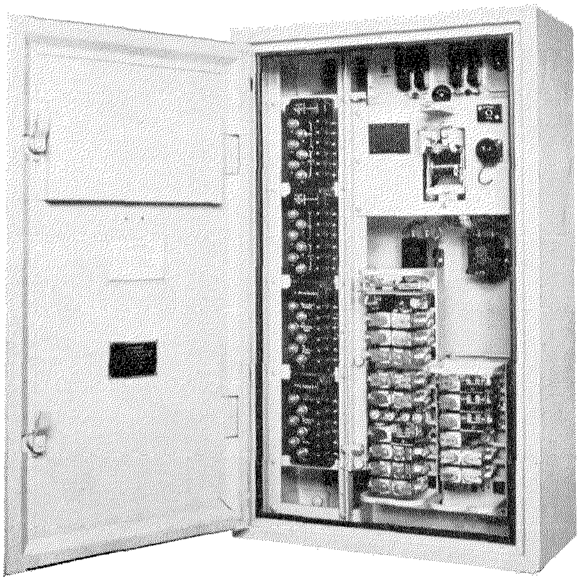

Fig. 7—Apparatus cubicle of wall-mounting type for substations, carrying moulded terminal blocks of kind shown in Fig. 9.

With the extended use of centralized schemes for large networks there is an increasing demand for means to give summated totals of readings derived from several sources. In self-contained areas of small extent, as for instance power stations, and where the totals do not in turn require to be transmitted, this need can be satisfied by the existing mechanical summators operating from a number of individual impulsing watthour meters. For a network of wide extent, readings corresponding to the individual load readings can be transmitted over the supervisory pilots to the control centre, where the summation can then be readily effected.

In one method, the local total load representing the average load over an agreed short period is converted into a signal code indicative of quantity, and this is transmitted at regular intervals (say, every two minutes) via the supervisory signalling channel, to the central point. There the various totals can be not only indicated individually, but also counted on registers, which may constitute an electromechanical summator for display according to local needs.

2. *Equipment Maintenance*

Once resort has been had to the use of supervisory control, it is highly desirable that the equipment be kept in first-class condition and departures from standard mean adjustment are detected before they can result in operation failure. The extent to which routine testing is necessary depends on the frequency of operation of the equipment, the type of maintenance personnel available, and the local conditions, such as distances involved and the availability of transport.

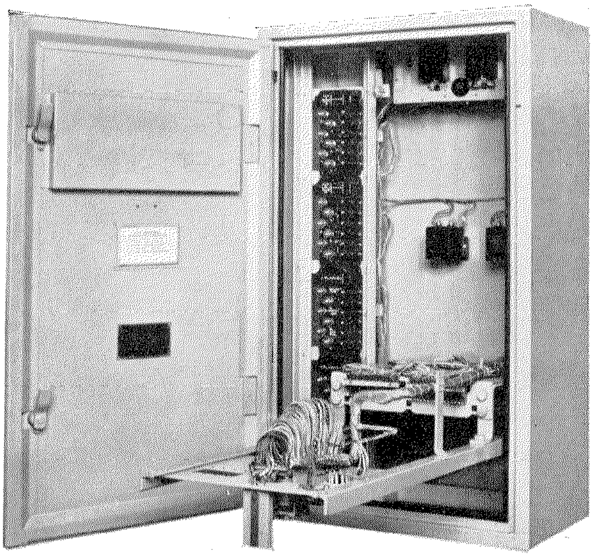

Fig. 8—Substation unit of the type shown in Fig. 7 without jack-in apparatus. The framework is hinged down to give access to the rear.

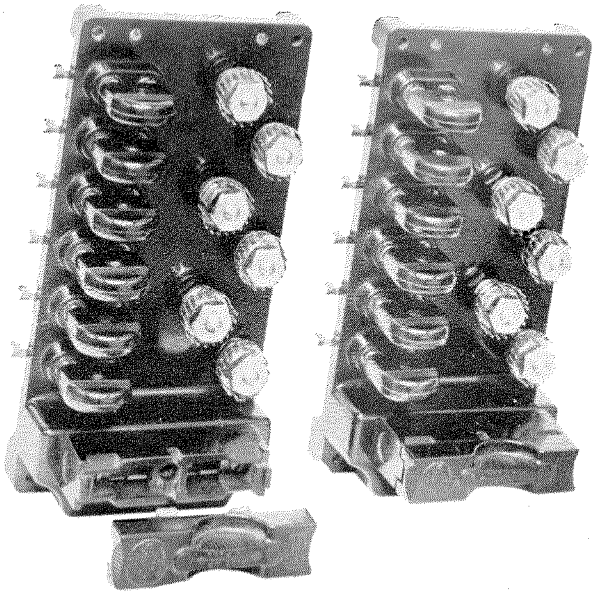

Fig. 9—Terminal block of the moulded type with isolating links. These blocks can incorporate signal lamps where necessary, or a fuse.

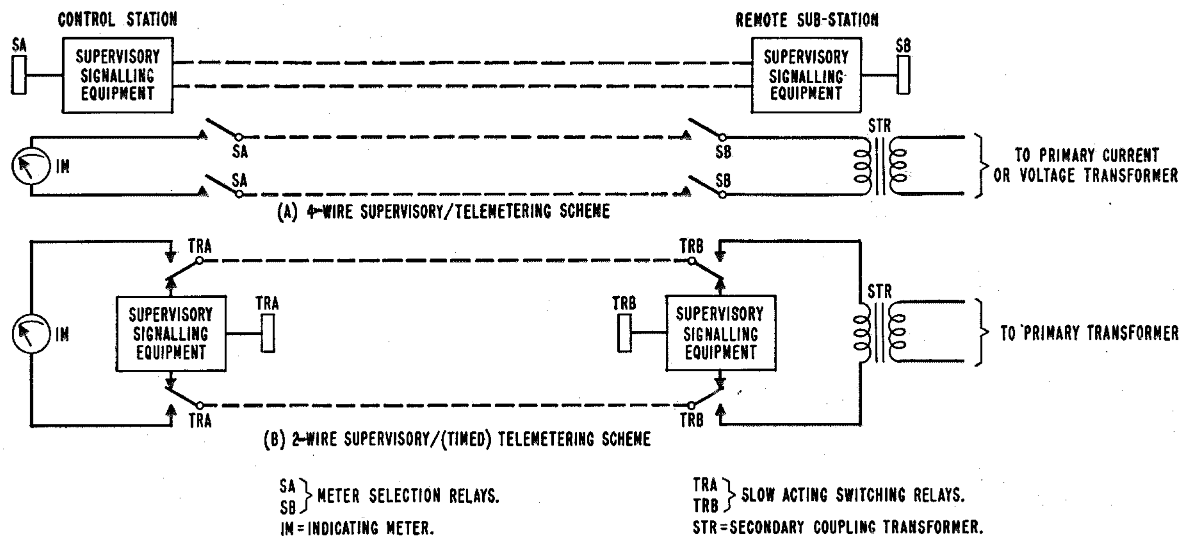
CONTROL STATION
REMOTE SUB-STATION
SA
SUPERVISORY SIGNALLING EQUIPMENT
SUPERVISORY SIGNALLING EQUIPMENT
SB
SA
SA
SB
SB
STR
IM
TO PRIMARY CURRENT OR VOLTAGE TRANSFORMER
(A) 4-WIRE SUPERVISORY/TELEMETERING SCHEME
TRA
TRB
IM
SUPERVISORY SIGNALLING EQUIPMENT
TRA
TRB
SUPERVISORY SIGNALLING EQUIPMENT
STR
TO PRIMARY TRANSFORMER
TRA
TRB
(B) 2-WIRE SUPERVISORY/(TIMED) TELEMETERING SCHEME
SA, SB } METER SELECTION RELAYS.
IM = INDICATING METER.
TRA, TRB } SLOW ACTING SWITCHING RELAYS.
STR = SECONDARY COUPLING TRANSFORMER.

Fig. 10—A direct metering scheme, which offers a ready solution for simple loading indications where the distances are short and a high degree of accuracy is not imperative.

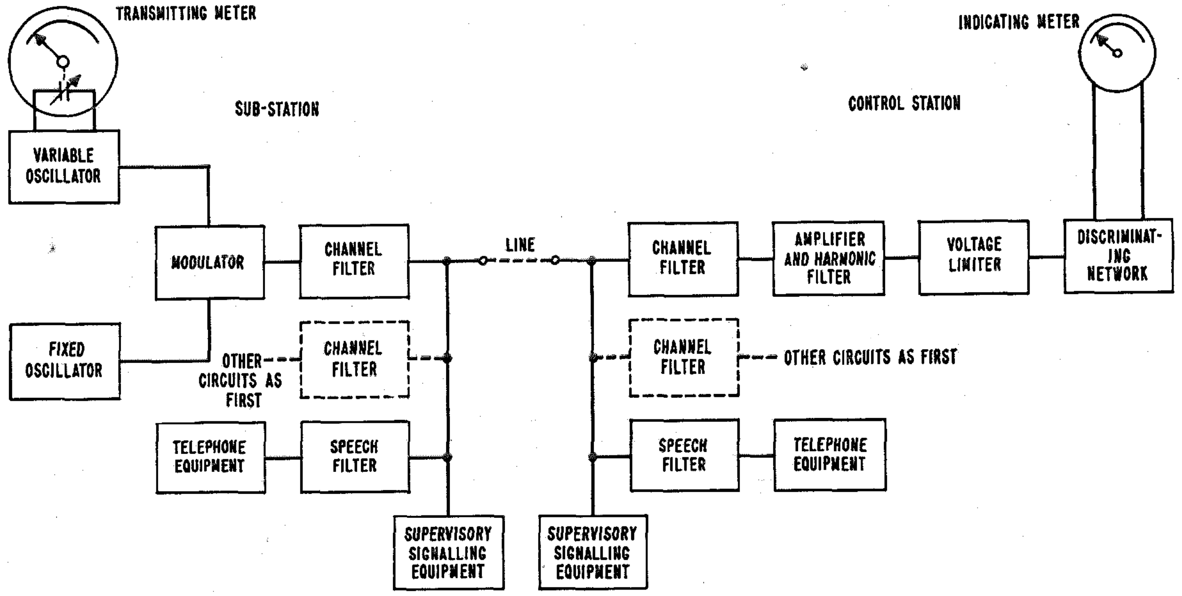

Fig. 11—Outline schematic of the voice-frequency continuous telemetering system.

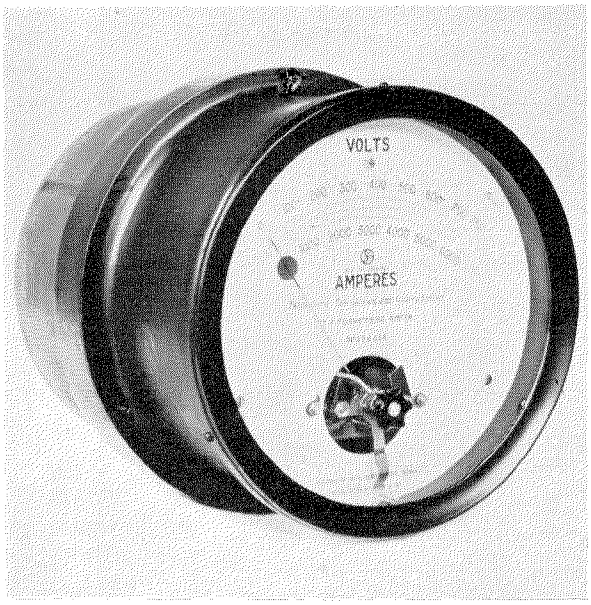

Fig. 12—Transmitting meter for voice-frequency continuous system has a small variable capacitor connected into the tuning circuit of a voice-frequency oscillator.

Equipment of the type described above depends primarily for its correct functioning on the common signal transmission and reception apparatus at the various stations, and for this reason the provision of portable routine test sets is most desirable. These are usually supplied in pairs, one being for use at the central control point, the other being available for use at any of the remote substations. By their use, it is possible to deal with practically all troubles single-handed, whereas without them most trouble-shooting becomes a two-man operation.

3. *Routine Testing Sets*

The set supplied for use at remote substations, for instance, is capable of simulating both the transmitting and receiving operations, which occur on the control-station equipment, i.e., it can transmit all signals that would normally be received at the substations and can check, too, the reception of all codes that can be transmitted from every substation.

Each set (see Fig. 13) is provided with a flexible connecting cord and multi-way plug, which for testing purposes is inserted in appropriate jacks fitted as standard equipment on each cubicle equipment.

The efficient operation of the equipment is very dependent upon reliable line conditions, and it is usual to include continuous line proving in which marked depreciation of line conditions is alarmed at the central point. In addition to this, the test sets normally include means for checking the line characteristics.

4. *Remote Testing Facilities*

It is sometimes argued that such testing as is indicated above, whilst fairly desirable, does not

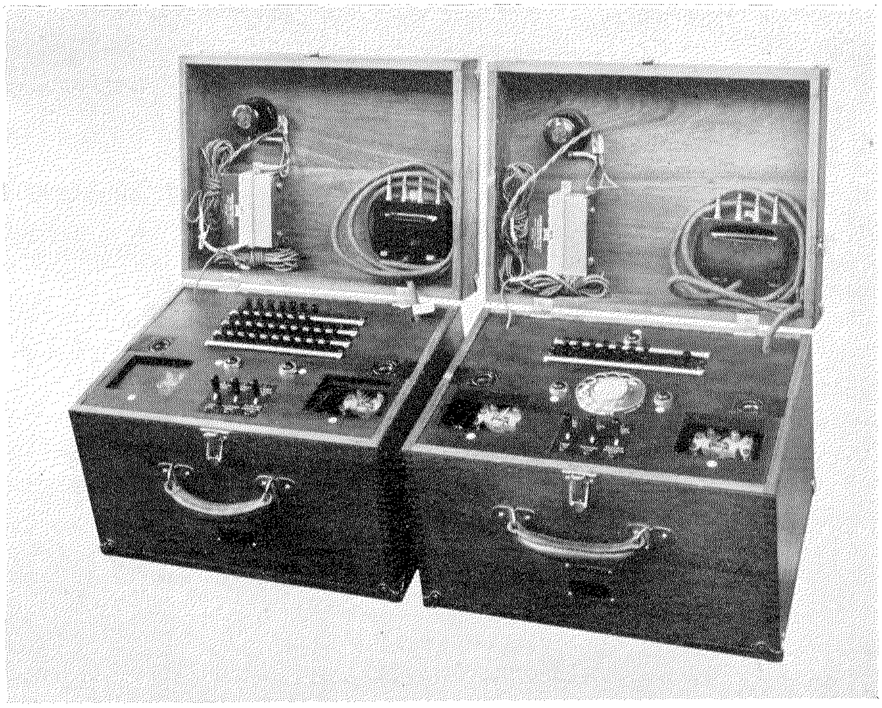

Fig. 13—Portable sets for routine tests are usually supplied in pairs, one for use at the central control point; other is for use at substation.

give ready means to the control attendant to check at intervals the through operation of the equipment, parts of which may not be called upon to function on some networks except at very infrequent intervals. To meet this criticism, some installations have included remote routine testing facilities. For each circuit breaker or similar device, subject to remote control and indication, a two-position relay of the latched-in type is provided on the substation control equipment. By operation of a common test key at the control point, the supervisory connections at the substation can thus be switched remotely, and for as long as may be required, from the switchgear to these relays. The latter can then be remotely controlled, just as if they were breakers, from the control room and the resultant indications checked there, thus bringing into operation all the associated control apparatus therein involved. On one such installation, the control engineer in this way routine tests some 14 remote substations once a week during a period in which switching operations are relatively quiet, and it is understood that this particular facility is viewed most favourably by the operating staff.

5. *Editor's Note*

Our overseas readers will be interested in a little more amplified description of the variable-frequency continuous (VFC) system to which Figs. 11 and 12 refer. The advantages claimed for the system are, in brief:

1. The band of frequencies used for one indication is narrow—50 to 80 cycles—and has sharply defined limits, so that a number of simultaneous indications can be transmitted over a single line or, alternatively, somewhat fewer if speech is required at the same time.
2. The indication is truly continuous at all parts of the scale, so that small changes are shown immediately.
3. There are no moving parts whatever other than the actual moving elements of the transmitting and indicating instruments.
4. The indication is independent of the line attenuation over a range of 0–30 decibels.
5. The indication is independent of cross-talk and other induced interference, provided the induced voltage is not greater than half the indicating voltage, and only then when the interference falls within the band employed.
6. The complete equipment may be run entirely from an alternating-current mains supply, the load being about 3 watts for the transmitting equipment and 10 watts for the receiving equipment. The accuracy of the indication is scarcely affected by fluctuations in the mains voltage, a variation at the transmitting station of ±10 per cent causing an error of only ±0.2 per cent in the indication. Variation of voltage at the receiving station has no appreciable effect.
7. The accuracy of the indication is only slightly affected by changing the oscillator valves. For instance, on test a maximum error of 0.5 per cent has been observed when the regular oscillator valves were replaced by very old ones. Any such error is easily checked and corrected at the transmitting station by adjustment of the oscillator. This is possible because of the presence of a positive signal at zero and full scale deflections. No zero adjustment is necessary on the actual meters other than those normally required.

Very-High-Frequency Triode Oscillator and Amplifier Circuits*

By GERARD LEHMANN

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PREVIOUSLY it has been shown[1] that triodes are capable of operating with substantially constant power gain and efficiency up to frequencies at which the dimensionless parameter φ reaches the value 2.5

$$\varphi = \frac{fd}{V^{1/2}},$$

where

f = frequency in megacycles per second,
d = cathode-anode distance in centimeters, and
V = supply voltage in volts.

Triodes still operate as oscillators, but with low efficiency, for values of φ exceeding 8.

These φ values correspond to frequencies much higher than those usually employed before the war. The inherent capabilities of triodes have been made useful through improvements in the associated circuits.

The replacement of the tube leads, in the form of thin rods passing through glass, by coaxial seals of large diameter and the use of resonant cavities have made possible an extension in frequency of two octaves in the application of triodes. The types of circuits that produce these results will be described and their operation and mode of adjustment will be analyzed.

• • •

1. General

Circuits associated with ultra-high-frequency tubes, such as Klystrons, have already been the object of numerous publications, so that it is not necessary to develop here all of the basic considerations.

The two principles that must be satisfied by the circuit (Fig. 1) are:

A. Circuits associated with electrodes that act on the electrons must be capable of being tuned to the desired frequency.

B. High voltage-magnification factors Q are almost always necessary.

These two conditions lead, as is well known in the case of Klystrons, to the construction of high-frequency circuits in the form of resonant cavities of revolution. Such cavities permit tuning at the frequencies contemplated and the symmetry of revolution results in high Q factors, everything else being equal.

In a cavity of revolution, the current density is constant along a circle having its plane at right angles to the axis of revolution, and it can be shown by the calculus of variations that this uniform distribution of current corresponds to minimum ohmic losses. Thus, cavity resonators fulfill the essential requirements for circuits associated with ultra-high-frequency triodes.

Tubes may be classified in two major groups according to the mode of flow of electrons: tubes having a plane structure as in Fig. 2, or a cylindrical structure as in Fig. 3.

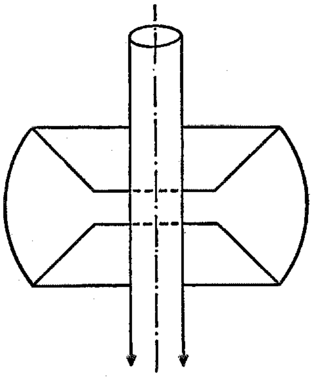

Fig. 1—Resonant circuit as used with ultra-high-frequency tubes.

Consider first the part played by the cavities in the case of a tube used as an amplifier and having negligible field penetration through the grid. This is the case for a very large amplification factor μ.

Omitting the power supply leads, the schematic is then very simple as may be seen in Fig. 4. On one side of the grid is the modulation or cathode cavity to which the energy for exciting the tube is applied, and on the other side the output or anode cavity to which the load impedance is coupled.

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* Reprinted from *L'Onde Electrique*, v. 26, pp. 357–366; October, 1946.

[1] Gerard Lehmann, "Application de l'analyse dimensionnelle aux tubes à très haute frequence, cas de la triode," *L'Onde Electrique*, v. 26, pp. 175–187; May, 1946: Also, *Electrical Communication*, v. 24, pp. 391–405; September, 1947.

2. *Q Factors*

If the operating conditions of the tube proper[1] are defined, the *Q* factors under load are determined and may be evaluated.

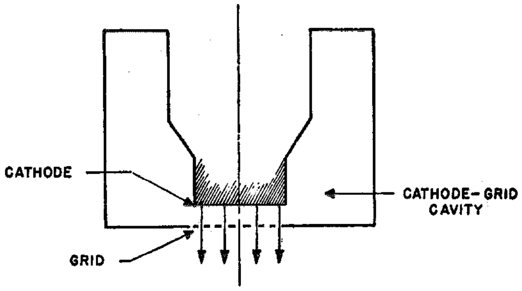

Fig. 2—Cavity of revolution associated with a tube having a plane cathode-grid structure.

To avoid possible confusion, it is desirable to define more precisely *Q* factor under loaded and unloaded conditions.

The *Q* factor in the unloaded condition of the cavities is that which would be measured with the filament of the tube inoperative and with the couplings to external impedances eliminated. This coefficient takes into account only the ohmic and dielectric losses in the cavity.

To be absolutely precise, it is sometimes preferable to make such a measurement with the cathode at normal operating temperature, but with the grid biased so that no electrons are emitted.

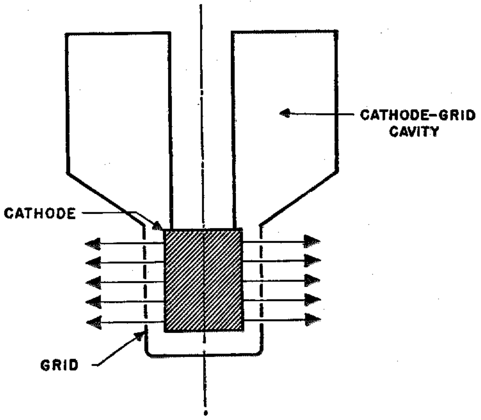

Fig. 3—Cavity of revolution associated with a tube having a cylindrical cathode-grid structure.

The unloaded *Q* factors may be calculated, theoretically, by classic methods when the geometric shape and the resistivity of the metals constituting the walls of the cavity are known.

In designing a circuit, it is extremely useful to make both the approximate theoretical calculation and the experimental measurement of unloaded *Q* factors of cavities, and not to accept differences of more than 10 to 20 percent between the two values. A larger difference indicates abnormal operation; most often an excessive resistance at certain contacts, which must be eliminated. Such an operation is of great help in disclosing weaknesses that very often would be much more difficult to locate with the tube in normal operation.

In contrast to the above, the *Q* factor under load does not depend critically on the losses in the circuits. Depending on the operating conditions of the tube, the load impedances are set by the characteristics of the tube, and the couplings between the cavities and the external circuits will be adjusted accordingly. In making these adjustments, the cavities are given *Q* factors that are determined primarily by the tube characteristics and not by the ohmic losses of the cavities.

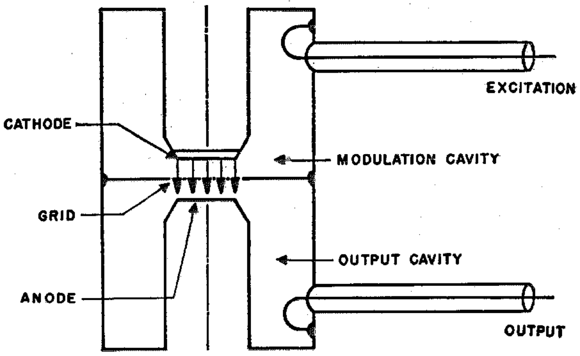

Fig. 4—Basic structure of a triode high-frequency amplifier using cavities of revolution.

Suppose two cavities, one of copper and the other of a higher-resistance material, are associated successively with the same tube. If the anode circuit is to present to the electron beam a load impedance imposed by the conditions of optimum operation, the load circuit must be strongly coupled to the copper cavity and weakly coupled to the resistive cavity. In both cases, the sum of the power used externally and the power lost inside of the cavity itself will be the same. Therefore, the Q under load is the same. The only difference is in the transfer efficiency between the circuit and the external load impedance, which is high in the case of copper and low in the second case.

Of course, it is indispensable to be able to effect an adjustment so that the Q coefficient in the idle condition may be greater than the desired loaded Q.

This is why it is possible to calculate the Q under load in the absence of information on the mechanical construction of the cavities.

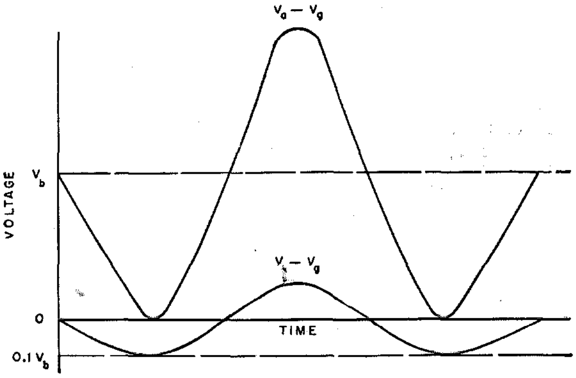

Fig. 5—General operating conditions for a triode transmitting tube. Grid potential O is plotted as the origin. Amplification factor μ is assumed to be infinite. V_a = anode voltage, V_g = grid voltage, and V_k = cathode voltage.

Let us make such an evaluation: In a previous article[1] concerned with the operation of triodes proper, it was shown that in general the following operating conditions (Fig. 5) are adopted for transmitting tubes:

A. Operation is class B.

B. Maximum potential difference between grid and cathode is 0.1 of the supply voltage.

C. Minimum potential difference between grid and anode should be zero volts.

All of the conditions of operation and the Q's are completely determined if we assume

$$\varphi = \frac{fd}{V^{1/2}}.$$

The tube operates in the quasi-steady state when φ is between 0 and 2.5, where
f = frequency in megacycles,
d = cathode-anode distance in centimeters, and
V = supply voltage in volts.

We now calculate the Q coefficients under load for this region of operation.

2.1 LOADED Q OF CATHODE CAVITY

2.1.1 *Active Current*

The peak current density is

$$A_{\text{peak}} = 2 \cdot 10^{-6} \frac{(0.1\,V)^{3/2}}{(d/4)^2},$$

approximately, in amperes per square centimeter.

This equation is obtained[1] by applying Langmuir's formula and taking into account the maximum grid voltage of $0.1\,V$, and the cathode-grid distance $d/4$. Therefore,

$$A_{\text{peak}} = 10^{-6} \frac{V^{3/2}}{d^2},$$

approximately, whence the fundamental component of the active current is

$$A_{\text{osc}} = \frac{1}{2} A_{\text{peak}} = 0.5 \cdot 10^{-6} \frac{V^{3/2}}{d^2}.$$

2.1.2 *Reactive Current*

As a first approximation, the density of the reactive current is the value of the current flowing through the capacitor formed by 1 square centimeter each of the cathode and grid surfaces under the effect of the cathode-grid potential difference.

$$A' = 0.1\,V \cdot \omega C_{kg} = 0.1\,V \cdot 2\pi f \frac{10^{-6}}{4\pi d/4}$$
$$= 0.2 \cdot 10^{-6} \frac{Vf}{d},$$

whence

$$Q = \frac{A'}{A_{\text{osc}}} = 0.4 \frac{fd}{V^{1/2}}$$
$$= 0.4\varphi.$$

Therefore, Q varies between 0 and 1 in the region of quasi-steady-state operation of the tube.

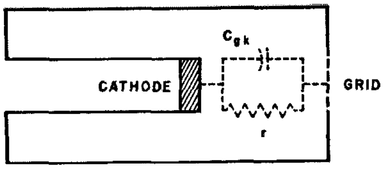

Fig. 6—Loading of the cathode-grid cavity.

Thus, the loading in the cathode-grid space of the tube appears as a resistance in parallel with a capacitance, the whole being shunted across the cathode cavity as indicated in Fig. 6.

In calculating the value of r, let s be the average transconductance of the tube in the operating region; for class-B operation, the active current is half of the peak current, and

$$r = \frac{0.1V}{0.5s \cdot 0.1V} = \frac{2}{s}.$$

In practice, s varies between 10 and 200 milliamperes per volt; therefore, r lies between 200 and 10 ohms.

It is possible to build coaxial lines having the above values of r as characteristic impedance. If a coaxial line, having a characteristic impedance

$$Z_0 = \frac{2}{s},$$

is directly terminated in the cathode-grid space, the coefficient of reflection will remain low and will be between 0 and 1 in the quasi-steady state (Fig. 7).

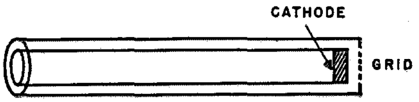

Fig. 7—Cathode space formed of a coaxial line terminated in its characteristic impedance: $Z_0 = 2/s$.

Such an arrangement is sometimes used in practice and allows a triode to be excited without any mechanical modification of the circuit over a very wide frequency band (several hundred megacycles). This arrangement is effective only in the case where there is no interpenetration of the fields existing on both sides of the grid, i.e., for very large values of μ (from 30 to 100).

Nevertheless, the general aspect of the phenomena is always as indicated above, and it is possible to take advantage of the low values of Q and r in the cathode circuit to adapt it to operation over wide bands of frequencies.

2.2 LOADED Q FACTOR OF ANODE CAVITY

First, it is useful to note that, except for grid current, the electron beam induces identical currents in the cathode and anode spaces. In particular, the fundamental component of the high-frequency current of amplitude $A/2$ in class-B operation is the same for both circuits.

Then, if V_e and V_u are the high-frequency voltages existing at the terminals of the cathode and anode cavities (excitation and anode oscillation voltages, respectively), the power gain is

$$G = \frac{V_u}{V_e},$$

and for the load resistances of the two cavities,

$$\frac{R_u}{R_e} = \frac{V_u}{V_e}.$$

The value of G, which is theoretically 10 in the scope of the assumptions previously made, is smaller in practice due to the interpenetration of the fields through the grid. Practically, G is between 5 and 8.

The loaded Q of the anode cavity is calculated as previously, the active current being

$$A_{\text{osc}} = 0.5 \cdot 10^{-6} \frac{V^{3/2}}{d^2}.$$

The reactive current is

$$A' = V \cdot 2\pi f \frac{10^{-6}}{4\pi \frac{3}{4} d}$$

$$= \tfrac{2}{3} \cdot 10^{-6} \frac{Vf}{d},$$

whence

$$Q = \frac{A'}{A} = 1.33\varphi.$$

Taking into account the current drawn by the grid, $Q = 1.8\varphi$.

In the case of actual tubes, the metallic and dielectric masses produce a total real capacitance about four times the capacitance of the active surfaces of the tube, which leads to $Q = 7.2\varphi$; for $\varphi = 2.5$, we find $Q = 18$.

The load resistance is about 10 times the values found for the modulation cavity; it lies, therefore, between 100 and 2000 ohms.

These conditions make impossible a direct connection between the output space and a coaxial transmission line, and the use of a resonant cavity is unavoidable. The construction and the operation of this cavity call for special care.

We thus arrive at a conception of the circuits associated with the triode as:

A. An excitation circuit with low impedance and Q, generally easy to realize.

B. An output circuit with a high impedance and Q, more difficult to build.

Further, the effect of the interpenetration of the fields through the grid makes the above ideal situation more complicated, and does away, as we shall see later, with the beautiful simplicity of the excitation circuit.

These principles now being established, we shall examine in greater detail the circuits that can be used for actual oscillators and power amplifiers.

3. *Oscillators*

It is well known that a given type of triode can work at its highest frequencies as an oscillator. For this reason, the study of ultra-high-frequency self-oscillating circuits deserves some attention. The two main problems are:

A. Design of the anode resonator.

B. Design of the feedback circuit to maintain oscillations.

3.1 ANODE RESONATOR

The anode resonator must meet four main conditions.

A. It must be able to resonate at the desired frequency.

B. It must not be readily susceptible to changes in the mode of oscillation.

C. It must have a Q factor sufficient to insure good transfer efficiency.

D. It must not interfere with the cooling of the anode.

Condition *A* can always be met by resonating the cavity at a partial mode, $2\lambda/4$, $3\lambda/4$, etc., if we are dealing with a coaxial line. It is best to select a mode of oscillation as low as possible ($\lambda/2$ or $\lambda/4$) to avoid spurious response at a frequency lower than that desired and to reduce the copper losses. Practically, with tubes having a plane structure, an oscillation in the $\lambda/4$ mode can almost always be adopted. In this case, as may be seen in Fig. 8, cooling is particularly easy, the anode being accessible from outside of the resonant cavity. The coupling to the load is effected by a loop. The only problem is the insertion of a series capacitance to prevent the anode direct voltage from reaching the grid.

All sliding contacts are to be avoided if good efficiency is desired. The Q under load being generally higher than 20 at the larger values of φ, the realization of a Q in the unloaded condition of between 200 and 1000 calls for rigorous precautions at all metal-to-metal contacts.

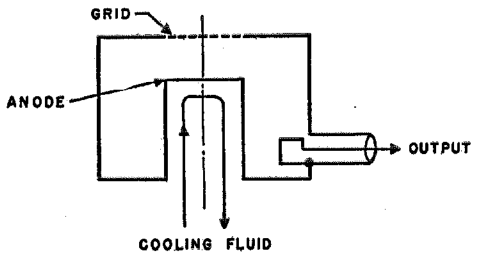

Fig. 8—Anode cavity resonator having $\lambda/4$ dimensions, for use with plane triode.

In the case of powerful tubes with cylindrical structure, the $\lambda/4$ circuit cannot generally be used for values of φ exceeding 2.5, the dimensions of the resonator becoming such that it has a tendency to disappear inside the tube.

A $\lambda/2$ resonator, shown in Fig. 9, can then be used successfully; with this circuit, no anomalous low-frequency oscillation can take place and stable operation is thus insured. On the other hand, use of the higher modes, $3\lambda/4$, $4\lambda/4$, etc., encourages mode changes and must be discarded for this reason.

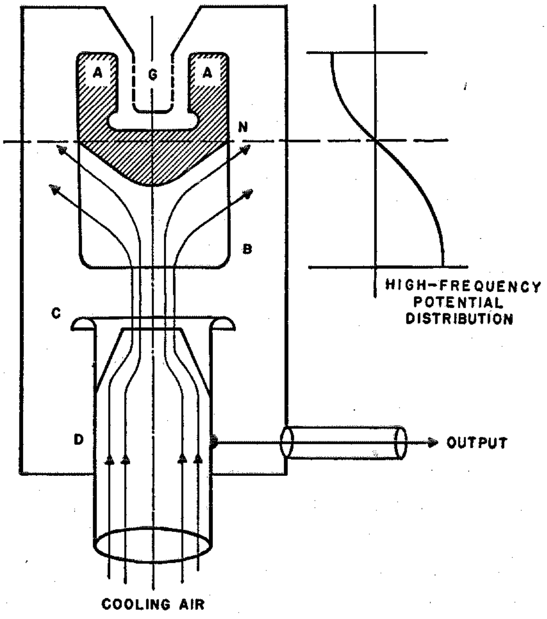

Fig. 9—A $\lambda/2$ resonator showing high-frequency potential distribution. The $\lambda/4$ coupling element *DC* is tubular to provide for cooling.

Fig. 9 shows the high-frequency potential at the anode A to be in phase opposition with the other end B of the resonating element. A potential node appears along circle N, which can be inside the bulb without causing any difficulties. Conductor NB may be a copper mass associated with the anode, serving as a heat exchanger with air circulation. The $\lambda/4$ element DC provides capacitive coupling of the load to surface B; cooling is accomplished by an air blast through the coupling element DC. An auxiliary device, described later, connects the high voltage to the anode.

Such a circuit has proved efficient for oscillators delivering high powers at very-high frequencies. It is particularly well adapted to the case of radar oscillators, where the fact that the plate is supplied with pulses of a few microseconds duration at a very high potential (20,000 volts) prohibits the use of a large bypass capacitor across the terminals of the supply source. The use of the half-wave resonator completely eliminates this difficulty.

This device was used as early as 1940 at the laboratories of Le Matériel Téléphonique by E. Touraton. Oscillators delivering a peak power of 600 kilowatts on a wavelength of 50 centimeters have been constructed more recently.

3.2 FEEDBACK CIRCUITS

The purpose of this circuit is to obtain from the anode cavity the power necessary for tube excitation, and to apply this power to the cathode-grid space in the form of a high-frequency voltage having the appropriate amplitude and phase for the maintenance of oscillations. In other words, the feedback circuit is a four-terminal network, two terminals of which are connected between the anode and the anode face of the grid, the other two terminals being connected between the cathode face of the grid and the cathode. For a given frequency, this four-terminal network is equivalent to a transfer impedance completely characterized by parameters X and Y, of the form $Z=X+jY$.

To say that the feedback circuit is adjusted to its optimum characteristics amounts to saying that X and Y have been given two particular values, X_m and Y_m.

For instance, for very low values of φ (negligible transit angle), the potential difference V_{gk} between the grid and the cathode, and the potential difference V_{ga} between the grid and anode, must be in phase agreement, and should also satisfy the relation

$$V_{gk}=0.1\,V_{ga}.$$

For other values of φ, the ratio $\alpha=V_{gk}/V_{ga}$ assumes complex (but perfectly defined) values for the fulfillment of the initial hypotheses.

The modulus of ratio α remains close to 0.1 for the quasi-steady states, and the value of its argument must be such that the voltage V_{ga} be exactly opposite in phase to the fundamental component of the current induced in cavity GA by the electron flow modulated by voltage V_{gk}. Thus, the conditions of Fig. 5 are satisfied in the operation of the oscillator and maximum efficiency is obtained.

At very high frequencies, the four-terminal feedback network is generally too complex to allow a calculation of the values of its elements as functions of α, the more so since the value of the ratio α itself is not well determined theoretically. The practical problem is then the design of a circuit capable of being experimentally adjusted so that the desired efficiency is actually obtained, with the certainty that this efficiency is the highest that can be delivered by the tube, independent of the circuit, under the imposed conditions.

In other words, we wish to be certain that the results given by the oscillator are limited by the electronic properties of the tube and not by poor design of the feedback circuit. That the feedback circuit depends on only the two parameters X and Y is fundamental, and provides the solution to the problem for a given frequency.

Thus, a feedback circuit may be set up in which two elements P and Q are experimentally adjustable and meet the two conditions

A. P and Q actually act as two *independent* parameters on the transfer impedance.

B. The optimum impedance Z_m be obtained for realizable values of P_m and Q_m.

The following method makes it possible to check whether these conditions can be fulfilled, and to find the optimum adjustments of P_m and Q_m. The oscillator is supplied from a constant-power source, and the high-frequency power delivered by the circuit is measured. By varying Q as a parameter and plotting the curves of useful power

$$W_u = f(P, Q),$$

a family of curves is obtained having the appearance of Fig. 10.

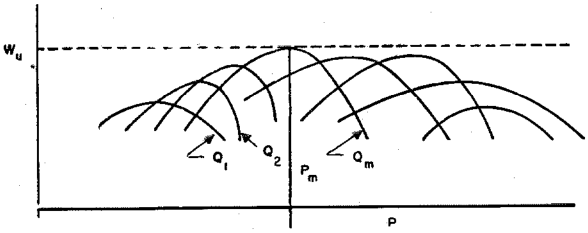

Fig. 10—Experimental measurements to determine optimum adjustments in oscillator circuit. The broken line indicates maximum output power at P_m and Q_m.

By making such a family of curves, the required independence of parameters P and Q is shown. The existence of maxima on the various curves of Q_1, Q_2, etc., and a maximum peak shows that the optimum values P_m and Q_m are within the range of adjustment. Under these conditions, the adjustment P_m, Q_m constitutes the optimum adjustment, and the broken line corresponds to the maximum power and efficiency of the tube.

For a given frequency, the same maximum efficiency $\eta_{\max}$ is obtained regardless of the type of feedback circuit adopted. However, feedback systems having appreciable ohmic losses give poorer results.

Two types of feedback circuits will be examined. The first is to be particularly recommended; the second one is defective, although often used.

3.3 Grid-Bell Feedback Circuit

The unit consisting of the tube and cathode and anode circuits has circular symmetry about an axis. It is preferable to adopt a feedback circuit also having a circular symmetry about the same axis so as to minimize ohmic losses. The necessity for having two independent adjustments on the feedback circuit leads to the arrangement of Fig. 11. The cathode and anode terminate in two coaxial lines of adjustable lengths, and the grid G, instead of completely separating the two cavities, is connected to an open tube, of length P, inserted inside the cathode cavity.

Generally, this open-ended insulated conductor is called the *grid bell.* Current flowing over the inner and outer surfaces of this bell will transfer energy between the anode and cathode, thus establishing the desired feedback. P and Q constitute the two parameters necessary for adjustment, and their values may be set experimentally without the slightest difficulty.

Thus, it is apparent that this oscillator may be operated under optimum conditions at a given frequency, as the *four necessary* adjustments are provided.

A. Frequency adjustment, effected by the anode resonator.

B. Load-resistance adjustment, effected by the load coupling.

C. Feedback adjustment P.

D. Feedback adjustment Q.

An oscillator that does not have these four adjustments cannot be operated at optimum efficiency, and the resulting decrease in efficiency is attributable to the circuit and not to the tube.

The use of grid-bell oscillators of the type shown in Fig. 11, adjusted in accordance with the principles set forth here, has made it possible in a series of experiments on a large number of

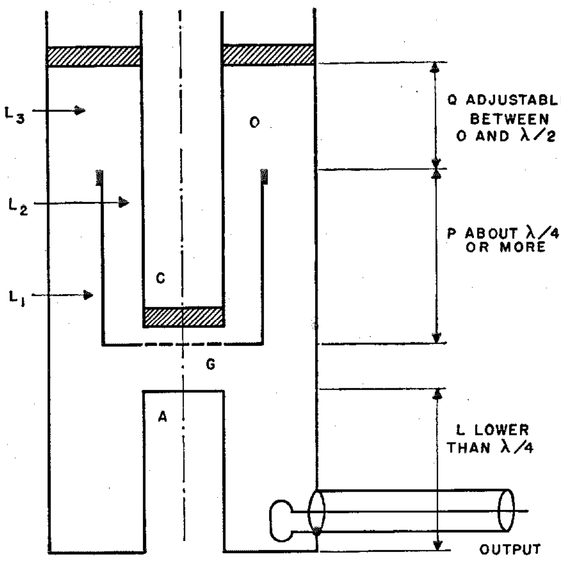

Fig. 11—Grid-bell oscillator construction. Feedback is provided through the coupling orifice O. P and Q are the two feedback adjustments.

tubes at frequencies between 300 and 3000 megacycles, to obtain results always equal to and often better than those obtainable with oscillators built on any other principle.

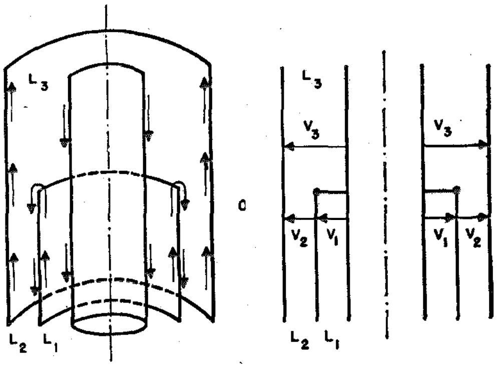

Fig. 12—Coupling at the junction of three coaxial lines, L_1, L_2, and L_3.

We shall now examine in more detail the operation of the feedback circuit: First, it is necessary to understand the mechanism of coupling at the junction of the three coaxial lines represented in Fig. 12. Near the orifice O, the total current flowing on each of the metal surfaces has exactly the same value; the directions of the currents are indicated by the arrows. Thus, at point O, the currents in the three coaxial lines L_1, L_2, and L_3 are identical. The integrals of the fields between the extreme conductors give the relation

$$V_3 = V_1 + V_2.$$

Fig. 13 shows the equivalent diagram, if each line is simulated at O by a two-terminal network. The three lines of Fig. 12 are thus *connected in series at point O.*

The feedback circuit of Fig. 11 can then be interpreted as follows: The lower end of line L_1 is connected in series with the grid-anode space; it passes the total current (convection and displacement currents) flowing through this space. After going through coaxial line L_1, the energy in the circuit flows through the reactance offered at O by line L_3, which is short-circuited at its upper end.

This energy then flows through line L_2 and finally induces the excitation voltage of the tube between cathode and grid. The adjustment of the length P of lines L_1 and L_2, and of the length Q of line L_3, are independent, permitting an accurate adjustment of the amplitude and phase of the excitation. Thus, we see that if $Q = \lambda/4$, the series reactance at O is infinite and the excitation is zero.

Lines L_1 and L_2 are of complex construction as they include parts of the glass structure of the tube, and it is difficult and useless to make a complete calculation of the system.

Length P is generally between $\lambda/4$ and $\lambda/2$, and length Q between 0 and $\lambda/2$.

The characteristic impedances for lines L_1, L_2, and L_3 may be chosen to fulfill additional conditions, such as operation over a wide frequency band. It is thus possible to build oscillator circuits capable of being tuned over a range of ±10 percent of the mean frequency by the tuning of only the anode circuit.

We still have to consider the direct-current-supply systems for the insulated conductors, such as the $\lambda/2$ resonator in Fig. 9, or the grid bell of Fig. 11. The problem consists of preventing any high-frequency-current leakage out of the system on the conductors supplying direct current.

The circuit of Fig. 12 gives a solution to this problem. Suppose we have to supply bias to a bell grid G (Fig. 14) through a wire F. Wire F is passed through two coaxial tubes, connected together at R through a piston, such that the length QR is equal to $\lambda/4$. This line being short-circuited $\lambda/4$ from point Q, offers at Q an infinite impedance. The result is that no high-frequency

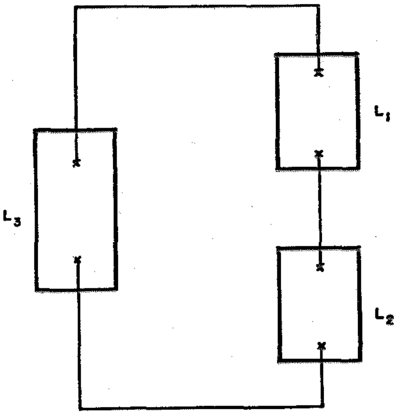

Fig. 13—The three lines of Fig. 12 are effectively in series.

current can flow along the supply wire *F*, which is in series at *P* with the infinite impedance *Q*.

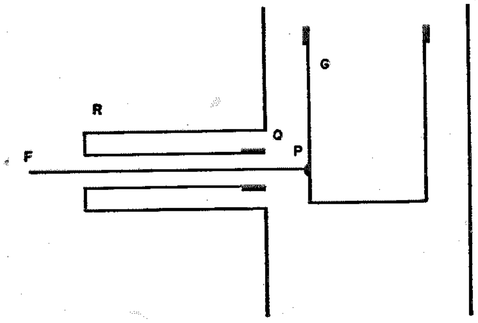

Fig. 14—High-frequency choke, permitting biasing of grid bell *G* through lead *F* without leakage of radio-frequency energy. This method avoids the need for a bypassing capacitor.

Such a system constitutes a high-frequency filter inserted in wire *F*. If wire *F* is suitably insulated, very high voltages can be applied at *P* without the necessity of a large bypassing capacitance (this is applicable to television and radar). Fig. 15 shows a complete radar oscillator built on these principles. On a wavelength of 50 centimeters, oscillators of this type allow the generation of powers of the order of 500 watts for continuous waves and 600 kilowatts peak power in pulse operation.

3.4 Oscillator With Feedback by Internal Capacitance

An oscillator circuit is often constructed by connecting the grid to the external shield of the circuits (Fig. 16), and adjusting the feedback with a cathode piston *P*. For feedback in such a circuit, reliance is placed on the existence of the cathode-anode capacitance due to the penetration of the fields through the grid. We then have only one variable for adjusting the feedback circuit, and the optimum adjustment cannot be obtained. Often, the system does not even start oscillating, the cathode-anode capacitance being too low. A remedy sometimes consists of placing inside the tube two rods connected to the cathode and entering the anode space through holes in the disk surrounding the grid. The cathode-anode capacitance is thus increased, and oscillations can be obtained. This system is not thought to be very advisable for two reasons:

A. The rods being in vacuum, capacitance *ka* is fixed, and the two necessary adjustment parameters are not available.

B. The tube thus built cannot be used in an amplifier, and, consequently, two types of tubes must be built; with and without an additional coupling capacitance. It is preferable to build only one tube having a cathode-anode capacitance as low as possible and to use a feedback circuit external to the tube.

4. *Power Amplifiers*

Fig. 4 represents a triode power amplifier in the ideal case where there is no interaction of the fields through the grid, i.e., when

$$c_{ka} = 0$$

and

$$\mu = \infty .$$

Unfortunately, this case is an ideal one, and the complications resulting from the existence of a capacitance between cathode and anode must be considered. To evaluate the effect of the latter, consider the electrostatic relation of a triode.

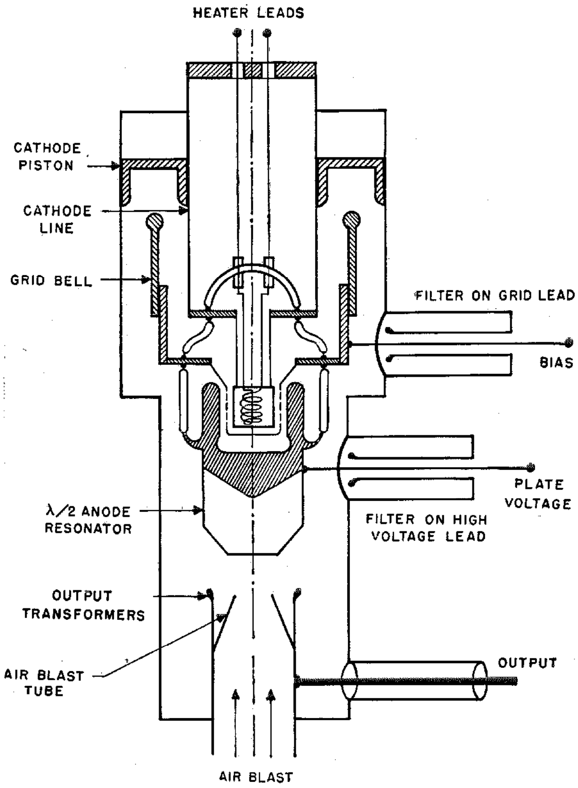

Fig. 15—Complete radar oscillator circuit delivering 500 watts continuous power and 600 kilowatts peak pulse power at 50 centimeters.

Let μ be the amplification coefficient, C_{kg} the cathode-grid capacitance, and C_{ka} the cathode-anode capacitance: We have

$$\frac{C_{ka}}{C_{kg}}=\frac{1}{\mu}.$$

For class-B operation, let I_1 be the fundamental component of the electron current flowing as a result of the excitation voltage V_e

$$I_1=V_e\frac{s}{2}.$$

The current brought to the cathode through capacitance C_{ka} is

$$I_2=V_u\cdot\omega C_{ka}=V_u\cdot\frac{1}{\mu}\omega C_{kg},$$

calling V_u the anode oscillating voltage; therefore,

$$\frac{I_2}{I_1}=\frac{V_u}{V_e}\cdot\frac{1}{\mu}\,\frac{\omega C_{kg}}{s/2}.$$

Now, $\omega C_{kg}/(s/2)$ is the Q under load of the excitation space, approximately 0.4φ. Hence, finally,

$$\frac{I_2}{I_1}=\frac{V_u}{V_e}\,\frac{0.4\varphi}{\mu}.$$

As V_u/V_e ranges from 5 to 10, when $\varphi=2.5$, we see that $I_2/I_1=1$ for $\mu=10$.

In this case, the anode voltage generates a cathode current equal to the excitation current.

This current, in turn, when flowing through the internal impedance of the source of excitation, causes disturbances in this circuit that make adjustment of the amplifier practically impossible. The current I_2 depends on V_u, i.e., on the adjustment of the anode circuit. In other words, the existence of capacitance C_{ka} has the effect of causing the tube excitation to depend on the adjustment of the anode circuit and, for $\varphi=2.5$, this effect cannot be tolerated; small variations of anode adjustment will make the amplifier oscillate (positive feedback) or, conversely, will suppress any amplification. Recourse must then be had to neutralization.

Neutralization of very-high-frequency coaxial amplifiers is relatively new and little has been published on the subject. Consequently, variations in both theory and practice may be expected.

In particular, at high frequencies, the transconductance of the tube and its internal impedance (when μ is low) become complex numbers. The effects of these imaginary terms have not yet been analyzed in a manner simple enough to allow a deduction of their effects on power amplifiers.

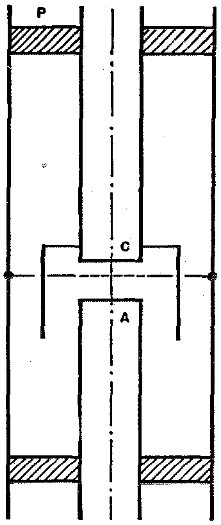

Fig. 16—Oscillator in which feedback is provided by the capacitance of two rods extending from the cathode through the grid disk.

An amplifier may be considered to be neutralized when it can be easily adjusted by methods usual for longer wavelengths to produce the maximum gain and efficiency predicted from the tube characteristics. In such a case, interaction between adjustments in the input and output circuits must be negligible. By varying the tuning of the output circuit, we must observe *simultaneously:*

A. Minimum plate current.

B. Maximum grid current.

C. Maximum useful power.

The accomplishment of these conditions has been studied in an amplifier delivering 1 kilowatt on a wavelength of 50 centimeters, with a tube having an amplification factor of 10. With the tube cold, it was found that the effect of capacitance C_{ka}, as expected, is to introduce a considerable interaction between the cathode and anode cavities.

The amplifier tube was inserted in a very-high-frequency coaxial bridge, which was adjusted (with the tube cold) to zero coupling between input and output circuits. Under these conditions, the reactances of the excitation and output circuits could be varied independently of each other. With the tube in operating condition, adjustments were made with the same ease and stability as at longer wavelengths.

An additional difficulty exists in neutralizing very-high-frequency amplifiers. The reactances of the leads to the electrodes inside the tube are important since these connections may belong simultaneously to one arm of the bridge and to one of the outside circuits. If, when the tube is cold, a strong oscillation is excited in the *anode* cavity by means of an auxiliary source, the tube is considered to be neutralized when the following two conditions are fulfilled:

A. No high-frequency potential difference exists between grid and cathode.

B. No current flows through the normal source of excitation for the tube.

To meet these two conditions, the neutralizing bridge must have two variables, while one is sufficient at low frequencies.

Fig. 17 represents the circuit used; R is the anode resonator; L_4 is the line supplying excitation energy to O' in the bridge circuit. The latter comprises the three lines, L_1, L_2, and L_3, introducing, between the spaces on both sides of the grid, the external coupling for neutralizing the effects of capacitance C_{ka}. As mentioned above, line element $O'C$ (including the input connection in the bulb of the tube) is common to the excitation circuit and to the neutralization bridge. The following reasoning accounts, in a simple manner, for the operation of the system.

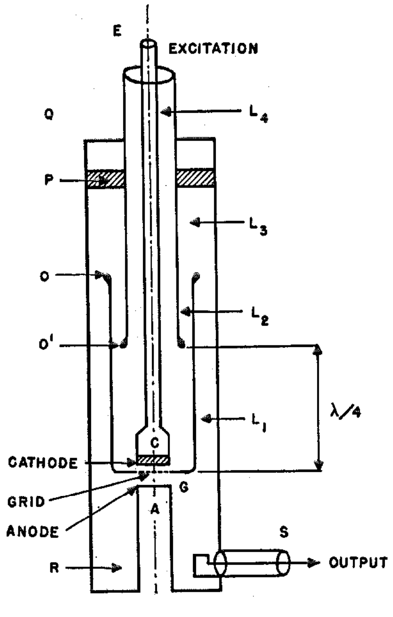

Fig. 17—Neutralized power amplifier.

The tube being inoperative (cold or strongly biased), let us excite cavity R by an auxiliary source connected to the output cable S. Due to the high Q factor of cavity R, considerable high-frequency voltage appears on the anode. If the excitation line E is open, no energy other than the small ohmic loss is dissipated on all the lines L_1, L_2, and L_3, coupled to the anode resonator. Thus a system of almost pure standing waves is established along these lines, and the nodes and loops are displaced when the line lengths are modified.

Neutralizing is obtained, subject to the two following conditions:

A. Cathode C is at a potential node; the grid-cathode potential difference is then zero, as desired.

B. Orifice O' is at a potential loop.

In such a case, point O' on line L_2 is a current node, and supplies no current to excitation line L_4 connected *in series* at point O'; the coupling between the source of excitation and the anode cavity is, therefore, zero. These conditions are obtained experimentally by adjustment of piston P and opening O'.

A detector placed at E will indicate the existence of condition B, and the tube itself, used as a diode (filament and grid), will indicate the existence of condition A. The operation is exactly like balancing an alternating-current bridge at low frequency. The Q's being high, it is almost indispensable to use an auxiliary source controlled by a quartz-crystal oscillator to obtain stable readings; naturally, this may be the driving stages of the transmitter.

The whole circuit lends itself to a quantitative analysis by writing successively the equations of the various coaxial lines and introducing the conditions discussed above. Without giving the calculations, we shall note that, in contrast with the case of oscillators, the positions of piston P and aperture O do not constitute two independent variables in the present case. A sequence of identical adjustments is obtained for various positions of O and P, associated in pairs. For this reason, it is mechanically convenient to make the final adjustment by moving only P. This is due to the fact that the feedback circuit carries only reactive energy.

There are several additional arbitrary variables, which are useful in satisfying secondary conditions, such as:

A. Termination at O' of excitation line L_4 in its characteristic impedance to eliminate standing waves on the line.

B. Maintenance of optimum conditions over a wide frequency band for television transmitters.

C. Introduction of controlled negative feedback to improve the linearity of amplification.

The bridge consists of three lines, which provide six variables (three lengths and three characteristic impedances). These offer great flexibility in the use of the system.

The use of this system has already made possible a solution of the difficult problem posed by the Columbia Broadcasting System for the transmission of color television. The transmitter, which is now undergoing a field test, comprises amplifiers of the type described using 6C22 tubes.[2]

The output stage delivers a peak power of 1 kilowatt on a 60-centimeter wavelength, and the video-frequency modulation extends to 10 megacycles; the total high-frequency bandwidth is 20 megacycles.

5. *Conclusion*

The developments of the last few years have resulted in great improvements in the field of meter wavelengths; then, under the influence of radar, in the field of centimeter wavelengths. The reduction in wavelengths has been so rapid that the field of the decimeter waves has been somewhat neglected.

We think that the wavelengths between 10 centimeters and 1 meter will have important applications, and this is the reason why interest is shown in the technique of triodes in special and well-adapted circuits.

The possibilities of generating continuous-wave power with triodes are roughly as follows:

A few deciwatts at 10 centimeters.
A few watts at 20 centimeters.
One kilowatt at 50 centimeters.
100 kilowatts at 3 meters.

We may expect in the future, and if the demand justifies the large work of elaboration that remains to be done, the appearance of tetrodes and pentodes for very-high frequencies, which will have powers and gains greater than those of triodes. The circuit technique may remain the same, comprising cavities of revolution for resonators and interwoven coaxial lines for feedback and neutralization circuits.

The Resnatron,[3] which is a tetrode tube delivering 50 kilowatts continuously on a 50-centimeter wavelength with an efficiency of 60 percent, is a landmark on the road to the future.

The expanding applications of transmitting tubes and circuits have always been a dominant factor in the development of radio and we do not yet see any sign that this trend has changed.

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Linear Theory of Bridge and Ring Modulator Circuits

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THE operation of single-sideband carrier telephone systems is based on frequency translation by copper-oxide modulators associated with filters. The impedance match between filters and modulators and the determination of modulator losses are, thus, technical problems of foremost importance. Since satisfactory operation of modulators is based on a linear relation between output and input voltages, a linear theory must be sufficient to solve these problems.

The principles of a general linear theory of modulator circuits have been developed by Caruthers and Kruse. They assume that the amplitude of the carrier voltage is much larger than the signal amplitude, the operation of the rectifiers then being controlled by the carrier voltage alone, except for negligibly short intervals when this voltage is near zero. This linear theory permits calculation of the performance of modulators (which may be considered as variable linear networks) from the rectifier characteristics, but it is far too intricate for practical design purposes. In the present paper, a more workable approximate theory is developed and some important ideal cases are discussed. Various circuit conditions existing in practical modulators have been analyzed separately; care must be taken in applying the results to a real circuit where several effects may interact. It is believed, however, that the present study will be of use to the designer.

* * *

1. *Introduction*

In papers dealing with rectifier-type modulators, various approximations have been proposed. In the most rudimentary theory, rectifiers are supposed to have zero forward impedance and infinite reverse impedance. With such a theory, the *ideal bridge-type modulator* is compared to a periodically shorting shunt circuit, and the *ideal ring-type modulator* to a periodic phase inverter. This permits a rough calculation of the loss, but, obviously, gives no information regarding optimum values of terminating resistance. A solution of the problem of matching can only be obtained if finite values of forward and reverse impedances are taken into account. The simplest approximation is to consider these impedances to be constant and purely resistive.

The theory of such *resistive modulators* working between purely resistive terminations has been developed by Caruthers,[1] and Parmentier.[2] If the modulator is terminated by filters or other reactive networks, or if its parasitic capacitance is taken into account, the exact analysis of the entire circuit entails laborious calculations involving Fourier series, even in the simplest cases. On the other hand, filter terminations usually increase modulator efficiency, due to their high or low impedance at unwanted sideband frequencies, and an estimate of this effect is desirable. To simplify the calculation, a natural assumption is to postulate *ideal filters* having constant resistance throughout the transmitted frequency range and an impedance approaching zero or infinity for the eliminated bands. For the ring modulator, this has been done in some cases by Caruthers;[1] for the bridge modulator, blocking circuits have been proposed by Katchatouroff and Lalande. Sections 2 and 3 of this paper are devoted to the theory of resistive bridge and ring modulators working between various ideal filter terminations.

Section 4 is devoted to some experimental results on bridge modulators. In practical rectifiers, the reverse impedance is mainly capacitive, and the results predicted by a purely resistive theory cannot be expected to hold. But *relative results*, such as the effect of ideal filters on the performance of a modulator compared to the performance of the same modulator working between pure resistances, are in very close agreement with measured values.

A further verification of the resistive theory is possible if the parasitic capacitance of the rectifiers is compensated; this has been achieved by means of a circuit first conceived by Clavier and Denis. Measurements have been performed by artificially controlling the forward and reverse resistances of the rectifiers (by adding series or shunt resistance), and all results are in accordance with theoretical formulas if definite values are designated for the effective resistances. The effective value of the reverse resistance is comparable to the mean value of the differential resistance over a range of negative voltages, and the effective forward resistance can be directly measured.

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[1] R. S. Caruthers, "Copper-Oxide Modulators in Carrier Telephone Systems," *Bell System Technical Journal*, v. 18, pp. 315–337; April, 1939.

[2] M. Parmentier, "Modulation par coupure et par inversion," *Annales des Postes, Télégraphes et Téléphones*, pp. 773–803, September, 1937; and p. 1028, December, 1937.

Starting with a modulator having compensated capacitance, the effect of increasing this capacitance has been systematically investigated, and an approximate theoretical expression is given in Section 4.

2. *Theory of Resistive-Bridge-Type Modulator*

According to the approximation of the resistive theory discussed in Section 1, the bridge modulator depicted in Fig. 1 will be replaced by a shunt resistance $r(t)$, which is a discontinuous function of time; it takes the value r_s (nearly a short circuit) during a half period of the carrier voltage, and the value r_o (nearly an open circuit) during the second half period. The variable resistance $r(t)$ will be connected between a generator of internal impedance Z_1 and a receiving impedance Z_2, as shown in Fig. 2. Both impedances are constant resistances at certain frequencies and infinite impedances at some other frequencies.

The following notations are used:

F = carrier frequency,
f = frequency of the generator,
$F+f$ = frequency of the useful sideband,
$\omega = 2\pi f$,
$\Omega = 2\pi F$, and
$\bar{\omega} = \Omega + \omega = 2\pi(F+f)$.

The analytic expression of the variable resistance $r(t)$ can be written

$$r(t) = r_s \frac{1-u(t)}{2} + r_o \frac{1+u(t)}{2}, \qquad (1)$$

where $u(t)$ is a periodically discontinuous function of time, alternating between $+1$ and -1, and suddenly jumping from one value to the other at frequency F. Its Fourier expansion is

$$u(t) = \frac{4}{\pi}(\sin \Omega t + \tfrac{1}{3} \sin 3\Omega t + \tfrac{1}{5} \sin 5\Omega t + \ldots). \qquad (2)$$

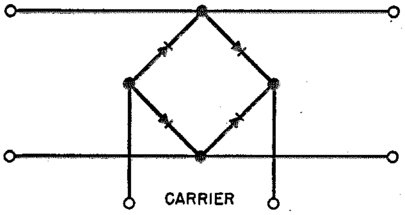

Fig. 1—Bridge-type modulator.

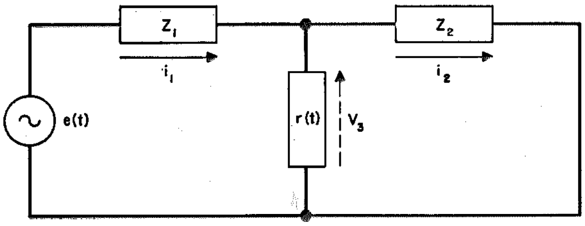

Fig. 2—Equivalent circuit of a bridge-type modulator with terminations.

2.1 OPERATION BETWEEN CONSTANT RESISTANCES

In the simplest case, both Z_1 and Z_2 are pure resistances, R_1 and R_2, respectively. If $e(t) = E \cos \omega t$ is the voltage of the generator, the output current (Fig. 2) is formally given by

$$i_2 = \frac{er}{(R_1+R_2)r + R_1R_2}, \qquad (3)$$

in which, for the sake of brevity, the functions of time have been intentionally omitted. This expression will now be transformed by replacing r by its value in (1) and by multiplying and dividing by a factor such that $u(t)$ is eliminated from the denominator. Since $u^2 = 1$, this factor is merely the conjugate of the denominator with respect to u, i.e., it is a similar expression containing $-u$ instead of u. After the above transformations, (3) becomes

$$i_2 = \frac{e}{2} \frac{(R_1R_2 + 2r_or_s)(R_1+R_2) + R_1R_2(r_o - r_s)u}{R_1^2R_2^2 + R_1R_2(R_1+R_2)(r_o+r_s) + (R_1+R_2)^2 r_or_s}.$$

The first term represents the input leak, whereas the second gives the modulation products. The two main products arise from the term containing $\frac{4}{\pi}E\cos\omega t\sin\Omega t=\frac{2}{\pi}E[\sin(\Omega+\omega)t-\sin(\Omega-\omega)t]$. If $\bar{\omega}=\omega+\Omega$ is retained as the only useful component, the corresponding output current will be

$$I_2\sin\bar{\omega}t$$

$$=\frac{E}{\pi}\frac{R_1R_2(r_o-r_s)\sin\bar{\omega}t}{R_1^2R_2^2+R_1R_2(R_1+R_2)(r_o+r_s)+(R_1+R_2)^2r_or_s}.$$

The loss in nepers, defined as half the logarithmic ratio[3] of the maximum available power $E^2/4R_1$ to the received power $R_2I_2^2$ in the useful sideband, is

$$A=\tfrac{1}{2}\log\frac{E^2}{4R_1R_2I_2^2}$$

$$=\log\pi+$$

$$\log\frac{R_1R_2+(R_1+R_2)(r_o+r_s)+\dfrac{(R_1+R_2)^2}{R_1R_2}r_or_s}{2(r_o-r_s)(R_1R_2)^{1/2}}. \quad (4)$$

Since this expression is symmetrical with respect to R_1 and R_2, the terminating resistances have a common optimum value, which may be found by differentiation:

$$R_1=R_2=2(r_or_s)^{1/2}=2\rho, \quad (5)$$

where ρ is the "mean resistance" of the rectifier.

The minimum insertion loss is then

$$A=\log\pi+\log\frac{\sqrt{r_o}+\sqrt{r_s}}{\sqrt{r_o}-\sqrt{r_s}}$$

$$=\log\pi+2\tanh^{-1}\left(\frac{r_s}{r_o}\right)^{1/2}. \quad (6)$$

The first term, which is equal to 1.15 nepers or 9.94 decibels, arises from the fact that only one modulation product is taken into consideration, whereas the second term gives the dissipative loss of the modulator. Usually, this last term is a small quantity, and the hyperbolic function is nearly equal to its argument. As practical formulas, we thus get

$$\left.\begin{aligned}&R_1=R_2=2\rho,\text{ where }\rho=(r_or_s)^{1/2},\\&\text{and}\\&A=9.94+17.4\,\alpha\text{ decibels, where }\alpha=(r_s/r_o)^{1/2}.\end{aligned}\right\} \quad (7)$$

It may be important to point out that optimum terminating conditions cannot be exactly described as matching some definite image impedances, for in the general case $R_1\neq R_2$ and expression (4) cannot be split into terms corresponding to reflection and interaction losses.

2.2 OPERATION WITH AN IDEAL FILTER

As a second case, consider a bridge modulator connected between a generator with constant ohmic resistance R_1 and a receiving impedance Z_2 that behaves as a pure resistance R_2 at the useful output frequency $f+F$ and presents an infinite impedance at all other frequencies. This is approximately obtained by the insertion between the modulator and the pure, constant, receiving resistance R_2 of a band-pass filter having a constant-k-type midseries impedance on the modulator side and a sufficiently narrow bandwidth to secure the elimination of all significant modulation products with the exception of the useful one, $F+f$.

By Thévenin's theorem, the generator, in connection with the modulator, is equivalent to a fictitious generator $e'(t)$ with an internal resistance $r'(t)$; both functions of time are easily computed by formal equations in terms of the u function, as in the preceding chapter:

$$\left.\begin{aligned}e'(t)&=e(t)\left(\frac{r_o}{R_1+r_o}\frac{1+u}{2}+\frac{r_s}{R_1+r_s}\frac{1-u}{2}\right),\\r'(t)&=R_1\left(\frac{r_o}{R_1+r_o}\frac{1+u}{2}+\frac{r_s}{R_1+r_s}\frac{1-u}{2}\right).\end{aligned}\right\} \quad (8)$$

Since $f+F$ is the frequency of output current $i_2(t)$, it will be determined by the equation

$$f+F\text{ component of }e'-r'i_2=R_2i_2. \quad (9)$$

If $e(t)=E\cos\omega t$, i_2 is purely sinusoidal; $I_2\sin\bar{\omega}t$, due to the ideal filter, and (9) can be written explicitly in terms of amplitudes.

$$\frac{E}{\pi}\left(\frac{r_o}{R_1+r_o}-\frac{r_s}{R_1+r_s}\right)-\frac{I_2R_2}{2}\left(\frac{r_o}{R_1+r_o}+\frac{r_s}{R_1+r_s}\right)=R_2I_2. \quad (10)$$

This gives immediately the amplitude of the output current. Since we are mainly interested in determining the optimum terminating resistances, we can now introduce the condition

$$R_2=\frac{R_1}{2}\left(\frac{r_o}{R_1+r_o}+\frac{r_s}{R_1+r_s}\right), \quad (11)$$

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[3] Natural logarithms are used throughout this paper.

which gives the optimum value of R_2 for a fixed value of R_1, and arises from (10) by the usual matching condition because the first member of this equation is independent of R_2.

With this condition, the output current and the loss can be expressed in terms of R_1 only:

$$\left.\begin{aligned} A &= \tfrac{1}{2}\log\frac{E^2}{4R_1R_2I_2^2} \\ &= \log\frac{\pi}{\sqrt{2}}+\tfrac{1}{2}\log\left(\frac{r_o}{r_o+R_1}+\frac{r_s}{r_s+R_1}\right) \\ &\qquad\times\left(\frac{r_o}{r_o+R_1}-\frac{r_s}{r_s+R_1}\right)^{-2}. \end{aligned}\right\} \tag{12}$$

The optimum value of R_1 is now found by equating the derivative of (12) with respect to R_1 to zero; this results in an equation of the third degree,

$$R_1^3-3r_or_sR_1-4\frac{r_o^2r_s^2}{r_o+r_s}=0. \tag{13}$$

Introducing notations $\rho=(r_or_s)^{1/2}$, $\alpha=(r_s/r_o)^{1/2}$, and $\cos\varphi=\tanh\,(2\tanh^{-1}\alpha)$, this equation becomes

$$\left(\frac{R_1}{\rho}\right)^3-3\frac{R_1}{\rho}-2\cos\varphi=0, \tag{14}$$

and its only positive real root is

$$R_1=2\rho\cos\frac{\varphi}{3}. \tag{15}$$

In practical cases, modulators are only slightly dissipative, and α is a small quantity. As a first approximation, $\cos\varphi=0$, $\varphi=\pi/2$, $\cos(\varphi/3)=\cos(\pi/6)=\sqrt{3/2}$, and $R_1=\rho\sqrt{3}=1.73\rho$. In the opposite case, for strongly dissipative modulators, $\alpha=1$, $\cos\varphi=1$, $\varphi=0$, $\cos(\varphi/3)=1$, and $R_1=2\rho$. It is easily shown that the optimum value of R_1 always lies between these extreme values.

Similar results are obtained for R_2 from (11), which can be rewritten

$$R_2=\frac{R_1}{(R_1/\rho)^2-1}, \tag{16}$$

and gives $R_2=\rho\sqrt{3}/2=0.87\rho$ in the case of small dissipation, and $R_2=2\rho/3=0.66\rho$ in the case of high dissipation.

For an ideal modulator terminated by optimum resistances, $\alpha=0$, the second term of (12) vanishes, and the loss of the modulator is $\log\pi/\sqrt{2}=0.8$ neper $=6.95$ decibels. To get a practical formula taking into account a small dissipative loss, the approximation must be carried one step further by expanding the first two terms of (15) and (16) into a series. This gives

$$\left.\begin{aligned} \frac{R_1}{\rho}&=\sqrt{3}+\tfrac{2}{3}\alpha, \\ \frac{R_2}{\rho}&=\tfrac{1}{2}\sqrt{3}-\tfrac{1}{3}(\sqrt{3}-1)\alpha, \\ A&=\log\frac{\pi}{\sqrt{2}}+\sqrt{3}\alpha. \end{aligned}\right\} \tag{17}$$

As practical formulas for slightly dissipative modulators, we thus have

$$\left.\begin{aligned} R_1&=1.73\rho, \\ R_2&=0.87\rho, \\ A&=6.94+15.02\alpha\text{ decibels.} \end{aligned}\right\} \tag{18}$$

The effect of the infinite output impedance on unnecessary products is thus a diminuation of the optimum impedances (more strongly at the filter side), and a reduction of 3 decibels in loss.

2.3 Input-Leak Blocking Circuit

In a bridge modulator circuit operating between constant resistances, the dominant part of the output current is the input leak, i.e., the component having the unmodified frequency of the generator. Consider, therefore, the effects of eliminating the input leak by means of a narrow-band elimination filter in the output circuit. When the frequency of the generator is very low compared to the carrier, this band-elimination filter is sufficiently approximated by a simple series capacitor.

The circuit will be idealized as follows:

A. The input impedance Z_1 is a constant resistance R_1.

B. The output impedance Z_2 is a constant resistance R_2, except at frequency f, where it is infinite.

The analysis begins as in the preceding paragraphs, but the output current i_2 may now contain all frequencies except f. Thus, (9) must be replaced by

$$e'-r'i_2=R_2i_2,$$

at any frequency except f. Introducing an unknown voltage v of frequency f, we can write at all frequencies including f,

$$e'-r'i_2=R_2i_2+v.$$

Solving formally for i_2 and equating its f component to zero, v is determined and i_2 can then be found. It is apparent that the amplitude I_2 of the $f+F$ component of i_2 is obtained from (10) as in the preceding circuit, and all the subsequent derivations are thus valid. An important fact is therefore brought to light, namely, that the 3-decibel gain over operation between constant resistances is already obtained by the elimination of the input leak. The elimination of other unnecessary products from the output circuit does not reduce the total power lost, but only produces a different distribution of the losses between the input and output circuits.

2.4 Operation Between Two Filters

We now suppose that Z_1 is a resistance R_1 at frequency f and is infinite for other frequencies. Also, Z_2 is a resistance R_2 at frequency $f+F$ and infinite for other frequencies.

The input and output currents are thus both purely harmonic; let us write,[4]

$$i_1=I_1\cos\omega t,$$

and

$$i_2=I_2\sin\bar{\omega}t.$$

The voltage v_3 across the modulator (Fig. 2) is

$$v_3=(i_1-i_2)r.$$

The f component of this voltage must be equal to $e-R_1i_1$ and the $f+F$ component to $-R_2i_2$. By introducing the amplitudes, we have

$$\left.\begin{aligned} E-R_1I_1&=\frac{r_o+r_s}{2}I_1-\frac{r_o-r_s}{\pi}I_2,\\ -R_2I_2&=\frac{r_o-r_s}{\pi}I_1-\frac{r_o+r_s}{2}I_2.\end{aligned}\right\}\qquad(19)$$

This set of equations is quite similar to the usual formulas for a symmetrical 4-pole network in terms of its self-impedances and transfer impedances.

The optimum terminating resistances, identical in this case to the image impedances, and the attenuation constant, can thus be deduced by conventional network theory. They are found to be

$$\left.\begin{aligned} R_1=R_2=R&=\frac{r_s+r_o}{2}\left[1-\frac{4}{\pi^2}\left(\frac{r_o-r_s}{r_o+r_s}\right)^2\right]^{1/2},\\ A&=\tanh^{-1}\left[1-\frac{4}{\pi^2}\left(\frac{r_o-r_s}{r_o+r_s}\right)^2\right]^{1/2}.\end{aligned}\right\}\qquad(20)$$

Introducing the notations $\rho=(r_sr_o)^{1/2}$, $A=2\tanh^{-1}\alpha$, $\alpha=(r_s/r_o)^{1/2}$, (20) may be written in a simpler form:

$$\left.\begin{aligned} \cosh A&=\frac{\pi}{2}\cosh A_o,\\ R\tanh A_o&=\rho\tanh A.\end{aligned}\right\}\qquad(21)$$

The first conclusion is that for an ideal switch ($r_s=0$, $r_o=\infty$) the image impedances become infinite and the attenuation constant becomes $\cosh^{-1}\pi/2$ = 1.02 nepers = 8.9 decibels. Since matching to an infinite impedance is impossible, the insertion loss will always be infinite. This is easily explained by the fact that operation of the modulator is then impossible, since the input and output currents must be equal during the open-circuit period of the switch, which condition cannot be presumed to hold for currents of different frequencies.

Thus, assuming a slightly dissipative modulator, we get as a second approximation

$$\left.\begin{aligned} R&=\frac{r_o}{2}\left(1-\frac{4}{\pi^2}\right)^{1/2},\\ A&=\cosh^{-1}\frac{\pi}{2}+\frac{\pi\alpha^2}{\left(\frac{\pi^2}{4}-1\right)^{1/2}},\end{aligned}\right\}\qquad(22)$$

leading to the practical formulas

$$\left.\begin{aligned} R_1=R_2&=0.39r_o,\\ A&=8.9+22.6\alpha^2\text{ decibels.}\end{aligned}\right\}\qquad(23)$$

It may be worth while pointing out that, for a strongly dissipative modulator, (21) are reduced to

$$\left.\begin{aligned} A&=\log\frac{\pi}{2}+A_o,\\ R&=\rho.\end{aligned}\right\}\qquad(24)$$

2.5 Operation Between Filter and Input-Leak-Blocking Circuit

From a practical point of view, the preceding circuit is inconvenient because of its high impedance and difficulty of control. In Section 2.3, it was noted that an input-leak-blocking circuit produces to a certain extent the same effect as a filter; we shall consequently try to replace the output filter of Section 2.4 with such a circuit.

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[4] It can be shown that a component like $I_2\cos\bar{\omega}t$ cannot be generated if the input current is to be of the form assumed.

Let us introduce the following hypotheses:

A. Z_1 is a pure resistance R_1 at frequency f, and infinite at other frequencies.

B. Z_2 is a constant resistance R_2 except at frequency f, where it is infinite.

The voltage $v_3(t)$ across the modulator is now simply divided between the output and the input impedances, the input taking the f component and the output the remainder. Thus,

$$r(i_1-i_2)=e-R_1i_1+R_2i_2.$$

Solving for i_2, replacing r by (1), and eliminating the u function from the denominator by means of conjugate multiplication, we get

3. *Theory of Resistive Ring Modulator*

With the approximations discussed in the introduction, the ring modulator represented in Fig. 3 is equivalent, during a half period of the carrier, to the lattice resistive network of Fig. 4, and during the next half period to a similar resistive network with r_o and r_s interchanged.

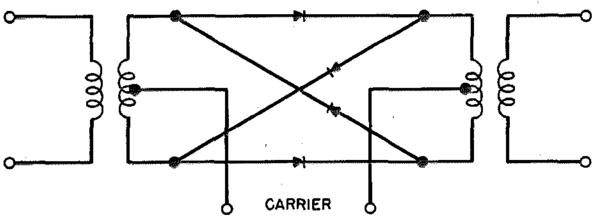

Fig. 3—Ring-type modulator.

The ring modulator may consequently be replaced by a fixed lattice resistive network followed by a periodic phase inverter. An ideal modulator ($r_s=0$, $r_o=\infty$) is reduced to the phase inverter alone. The ideal transformers shown in

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$$i_2=\frac{-e\left(\frac{r_o+r_s}{2}+R_2\right)+i_1\left[R_1R_2+(R_1+R_2)\frac{r_o+r_s}{2}+r_or_s\right]-[i_2(R_2-R_1)+e]\frac{r_o-r_s}{2}u}{r_or_s+\left(\frac{r_o+r_s}{2}\right)R_2+R_2^2}. \tag{25}$$

The first two terms are merely of frequency f, whereas the third term has no f component. Since i_2 cannot contain an f component, the first two terms cancel each other. Dropping these terms, solving for i_2, and taking the amplitude of the $f+F$ component, we get

$$I_2=\frac{E}{\pi}\frac{r_o-r_s}{R_1R_2+\frac{1}{2}(r_o+r_s)(R_1+R_2)+r_or_s}.$$

The loss is

$$\left.\begin{aligned}A&=\tfrac{1}{2}\log\frac{E^2}{4R_1R_2I_2^2}\\&=\log\frac{\pi}{2}+\log\frac{[R_1+\frac{1}{2}(r_s+r_o)][R_2+\frac{1}{2}(r_s+r_o)]-\frac{1}{4}(r_o-r_s)^2}{(r_o-r_s)(R_1R_2)^{1/2}}.\end{aligned}\right\} \tag{26}$$

Apart from the first constant term, this is identical with the formula for the insertion loss of a symmetrical 4-pole network having an image impedance $\rho=(r_or_s)^{1/2}$ and a transfer constant $A_o=2\tanh^{-1}\alpha$. Thus, (24), obtained in the preceding case for largely dissipative modulators, becomes

$$\left.\begin{aligned}R_1&=R_2=\rho,\\A&=3.92+17.4\alpha\text{ decibels.}\end{aligned}\right\} \tag{27}$$

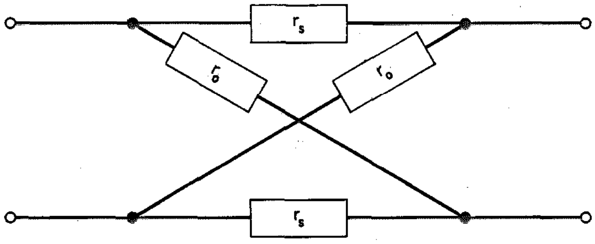

Fig. 4—Lattice-resistive network, equivalent, during one-half cycle of the carrier, to Fig. 3. During the second half cycle, r_o and r_s are interchanged.

Fig. 3 are only provided to supply the carrier voltage, and will be supposed to have 1:1 ratios.

Using the periodically discontinuous function $u(t)$, the equations of a phase inverter (Fig. 5) are obviously

$$\left.\begin{aligned} i_2(t) &= u(t)i_1(t),\\ v_1(t) &= u(t)v_2(t),\end{aligned}\right\} \qquad (28)$$

and are similar to the equations of an ideal transformer having a ratio u, which is a function of time.

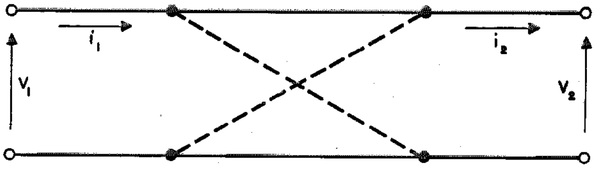

Fig. 5—Equivalent circuit of an ideal ring-type modulator.

If a phase inverter is connected between a generator of voltage $E\cos\omega t$ with internal resistance R and a matched receiving resistance R, the output voltage will be $\frac{1}{2}Eu(t)\cos\omega t$. Using the Fourier expansion (2) of $u(t)$, the output voltage appears to contain the angular frequencies $n\Omega\pm\omega$ (where n is odd). If only one of the main products, e.g., $\bar{\omega}=\Omega+\omega$, is considered as useful, the loss in nepers of the ideal modulator is easily found to be

$$\left.\begin{aligned} A &= \log\frac{\pi}{2}\\ &= 0.45 \text{ nepers}\\ &= 3.91 \text{ decibels.}\end{aligned}\right\} \qquad (29)$$

Consider now the lattice network of Fig. 4. Its image impedances are

$$\rho = (r_o r_s)^{1/2}, \qquad (30)$$

and its image attenuation in nepers is given by

$$\left.\begin{aligned} A_o &= 2\tanh^{-1}\alpha\\ &= 2\tanh^{-1}(r_s/r_o)^{1/2}.\end{aligned}\right\} \qquad (31)$$

Since α is usually a small quantity, A_o in decibels is approximately given by

$$A_o = 17.4\alpha. \qquad (32)$$

The optimum terminating impedances for a ring modulator working between pure resistances are thus $R_1=R_2=\rho$, and the attenuation for the sideband considered is

$$A = 3.92 + 17.4\alpha \text{ decibels.} \qquad (33)$$

Since a phase inverter may be interchanged with any 4-terminal resistive network, the preceding results obviously hold good in more general cases, provided that at least one of the terminating impedances is a pure and constant resistance.

The results are quite different for a ring modulator connected between two filters. Only the approximation of "ideal filters," having very low or very high impedance for unwanted frequencies, leads to simple formulas. Suppose that the modulator is working between impedances Z_1 and Z_2. We shall distinguish between the following cases:

$$\text{Case 1}\left\{\begin{array}{l} Z_1 = R_1 \text{ at the signal frequency } f\text{, and is infinite for other significant frequencies.}\\ Z_2 = R_2 \text{ at the useful sideband } f+F\text{, and is infinite for other significant frequencies.}\end{array}\right.$$

$$\text{Case 2}\left\{\begin{array}{l} Z_1 = \text{same as in Case 1.}\\ Z_2 = R_2 \text{ at the useful sideband } f+F\text{, and is zero for other significant frequencies.}\end{array}\right.$$

Case 1 is approximated by a modulator connected between two midseries-terminated filters. The opposite case 1′, where both filters have midshunt terminations, may be deduced from the first by means of the duality principle, and will therefore not be considered. In Case 2, the filters have inverse terminations. Case 2′, where Z_1 and Z_2 would be interchanged, is not distinct from Case 2, as is indicated by the reciprocity theorem.

The theory of Case 1 is in all respects similar to the theory developed for the bridge modulator between the same terminations (Section 2.4). The image impedance and attenuation are

$$\left.\begin{aligned} R_1 = R_2 &= \rho\frac{\tanh A}{\tanh A_o},\\ A &= \cosh^{-1}\left(\frac{\pi}{2}\cosh A_o\right).\end{aligned}\right\} \qquad (34)$$

For an ideal modulator ($A_o=0$), the minimum attenuation is $\cosh^{-1}(\pi/2)=1.02$ nepers, and the image impedances are infinite. This result has been found by Caruthers[1] and Kruse,[5] and is easily explained by the fact that the first equation in (28) cannot hold with currents i_1 and i_2 of different pure frequencies.

A similar result is obtained for Case 1′ with the same formula for A and the inverse impedances;

$$R_1=R_2=\rho\frac{\tanh A_o}{\tanh A}, \tag{35}$$

giving zero image impedance if $A_o=0$.

When $A_o=0$, a match is possible in both Cases 1 and 1′, but the attenuation as given by (7) is always larger than that which can be obtained between inverse terminations (Case 2). See Fig. 6.

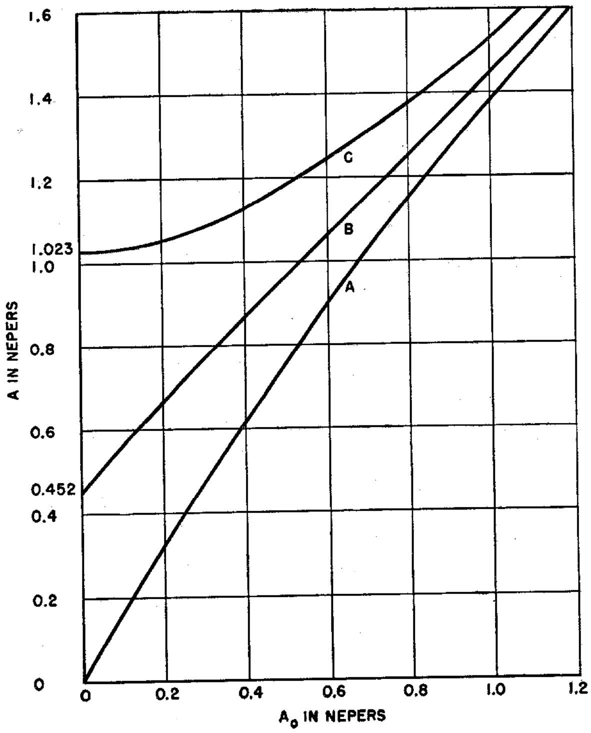

Fig. 6—Attenuation of a resistive-ring-type modulator:
A. between inverse ideal filters; $\sinh A=\frac{\pi}{2}\sinh A_o$.
B. between resistances; $e^A=\frac{\pi}{2}e^{A_o}$.
C. between similar ideal filters; $\cosh A=\frac{\pi}{2}\cosh A_o$, where $A_o=2\tanh^{-1}(r_s/r_o)^{1/2}$.

The analysis of Case 2 is easily carried out by writing the equations of a ring modulator in terms of the instantaneous currents and voltages. They are

$$\left.\begin{aligned} v_1(t)&=\rho\coth A_o i_1(t)-\frac{\rho}{\sinh A_o}u(t)i_2(t),\\ &\text{and}\\ v_2(t)&=\frac{\rho}{\sinh A_o}u(t)i_1(t)-\rho\coth A_o i_2(t),\end{aligned}\right\} \tag{36}$$

and are deduced from the usual equations of an attenuator having image parameters ρ and A_o by simply multiplying the transfer impedances by $u(t)$, which gives the periodic phase reversal. Due to the nature of the terminating impedances, the output voltage and the input current are purely harmonic:

$$v_2(t)=V_2\cos\omega t,\quad\text{and}\quad i_1(t)=I_1\sin\omega t.$$

Introducing these values in (36), $v_1(t)$ and $i_2(t)$ may be calculated. Since we are only interested in the amplitude V_1 of the f component of $v_1(t)$, and in the amplitude I_2 of the $F+f$ component of $i_2(t)$, they will be calculated using the expansion (2) of $u(t)$. Rearranging the results, the two relations between the amplitudes are

$$\left.\begin{aligned} V_1&=\frac{\rho}{\cosh A_o}\left(\frac{4}{\pi^2}\frac{1}{\sinh A_o}+\sinh A_o\right)I_1\\ &\qquad\qquad-\frac{2}{\pi}\frac{\rho}{\sinh A_o}I_2,\\ V_2&=\frac{2}{\pi}\frac{\rho}{\sinh A_o}I_1-\rho\,\mathrm{cotanh}\,A_o I_2.\end{aligned}\right\} \tag{37}$$

These are standard equations for a 4-terminal network; the image parameters are easily obtained by the usual network theory. They are found to be

$$\left.\begin{aligned} R_1&=\rho\tanh A_o\coth A,\\ R_2&=\rho\tanh A\coth A_o,\\ A&=\sinh^{-1}\left(\frac{\pi}{2}\sinh A_o\right).\end{aligned}\right\} \tag{38}$$

For an ideal modulator ($A_o=0$), A is also zero, and the ratio R_2/R_1 has the value $\pi^2/4$. This has been found by Caruthers. For a slightly dissipative modulator, (38) becomes approximately

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[5] S. Kruse, "Theory of Rectifier Modulators," Thesis; Stockholm, 1939.

$$\left.\begin{aligned} R_1&=\frac{2}{\pi}\rho,\\ R_2&=\frac{\pi}{2}\rho,\\ A&=\frac{\pi}{2}A_o \text{ neper or } 27.2\alpha \text{ decibels.}\end{aligned}\right\} \quad (39)$$

In Fig. 6, the attenuations of Cases 1 and 2 are compared; when A_o is large, (34) and (38) become

$$\left.\begin{aligned} A&=\log\frac{\pi}{2}+A_o,\\ R_1&=R_2=\rho.\end{aligned}\right\} \quad (40)$$

These are identical to the formulas dealing with modulators placed between pure resistances.

4. *Experimental Results*

4.1 EFFECT OF IDEAL FILTER TERMINATIONS

The transmission loss of a bridge modulator has been measured under different terminating conditions very closely approximating the assumptions of the theoretical analysis of Section 2. The same modulator, composed of 4 Westinghouse 3-millimeter copper-oxide rectifiers, has been used in all experiments. The carrier voltage was 1.2 volts at 90 kilocycles, i.e., 0.6 volt per rectifier; the signal voltage was 0.03 volt at 90.1 kilocycles. The difference frequency, 0.1 kilocycle, was taken as the useful sideband. The measuring circuit is shown in Fig. 7.

A first series of measurements (Fig. 8, Curve A) was carried out with purely resistive and symmetrical terminations. The minimum loss, 12.5 decibels, occurs when $R_1=R_2=1700$ ohms.

A second series of measurements was carried out with a coil of 80 millihenries inserted in series with the receiving resistance. This coil is practically a short circuit for the useful frequency, 0.1 kilocycle, and a very high impedance for the other frequencies (90 kilocycles and above). According to the theory of Section 2.2, the optimum terminating conditions in this case are such that $R_2=R_1/2$; see (18). To check this theoretical result, measurements were made with $R_2=R_1$ (Fig. 8, Curve B), and with $R_2=R_1/2$ (Fig. 8, Curve C). In accordance with the theory, the second condition gives a lower minimum attenuation, and the reduction of this minimum as compared to Curve A is actually 3 decibels. Furthermore, the difference between the attenuation of Curves B and C is of the order of the reflection loss obtained by terminating Z with $2Z$ (i.e., 0.56 decibels). According to (5) and (18), the optimum generator impedance in this case must be reduced to 87 percent of the optimum terminating resistance, which is 1700 ohms in the first case. Since $0.87\times1700=1479$ ohms, this is confirmed satisfactorily by experiment.

A final series of measurements was carried out by adding a capacitor of 0.1 microfarad in series with the generator. This presents a very high impedance at 0.1 kilocycle and a short circuit for all other important frequencies. The coil in series with the output impedance was retained. According to (27), the optimum terminating resistances must now be halved; the minimum attenuation, as compared with the first case, must be reduced by 6 decibels. This was confirmed by experiment (Fig. 8, Curve D).

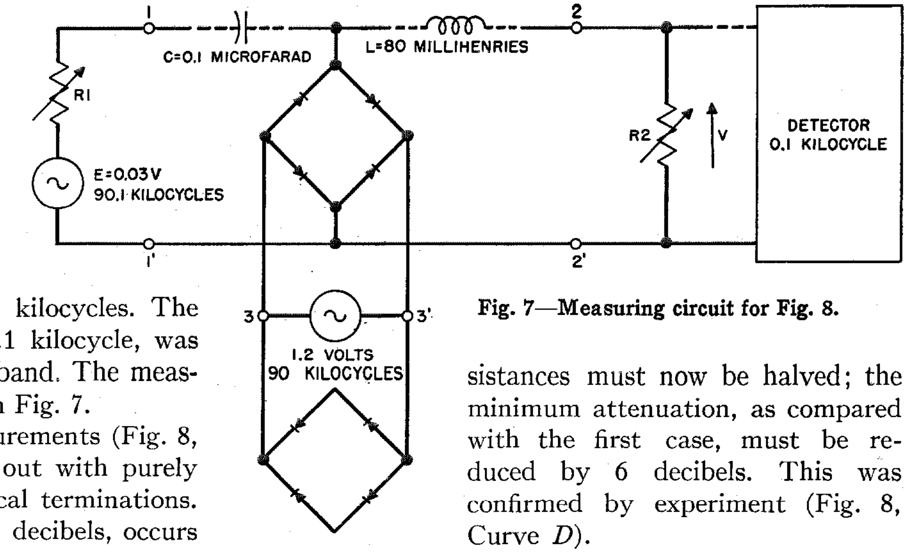


Fig. 7—Measuring circuit for Fig. 8.

4.2 COMPENSATION OF PARASITIC CAPACITANCE

The parasitic capacitance C of a rectifier can easily be deduced from an impedance measurement of two identical rectifiers in opposing parallel connection (Fig. 9). If alternating voltage be applied, each rectifier successively passes and blocks; the other rectifier is simultaneously

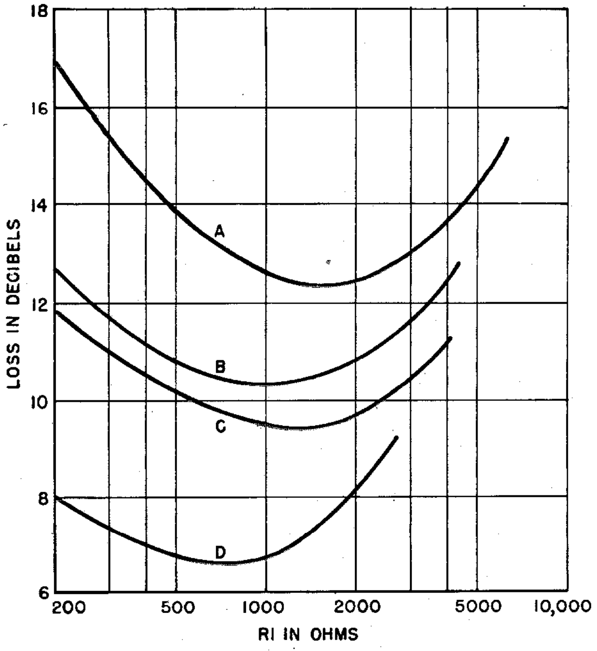


Fig. 8—Transmission loss of a bridge-type modulator between various terminations:

- **A. between pure resistances $R_1 = R_2$.**
- **B. with L inserted and $R_2 = R_1$.**
- **C. with L inserted and $R_2 = R_1/2$.**
- **D. with both L and C inserted and $R_2 = R_1$.**

in the opposite condition. Since the forward capacitance of a rectifier is small compared to its reverse capacitance, while the reverse resistance is negligible when shunting the forward resistance, the mean value of the reverse capacitance in shunt with the mean value of the forward resistance is obtained. This measurement was performed with the rectifier of the preceding experiment, and with the same voltage; 0.6 volt at 90 kilocycles. The measured values are $r_s = 140$

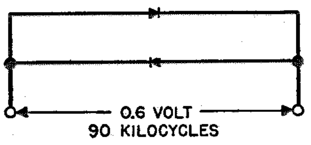


Fig. 9—Measurement of rectifier impedance.

ohms and $C = 1300$ micromicrofarads. This parasitic capacitance can be considered as shunting the bridge modulator.

An exact compensation of the parasitic capacitance at all frequencies $nF+f$ requires a shunt-reactance dipole including an infinite number of elements. A partial compensation is possible, however, if only the generator frequency f and the useful output frequency $F+f$ are considered. The circuit of Fig. 10, of the type first suggested

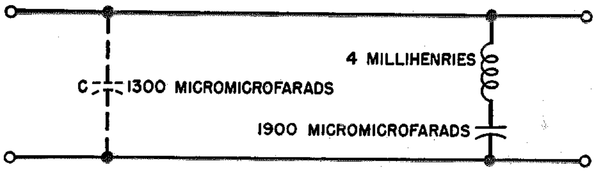
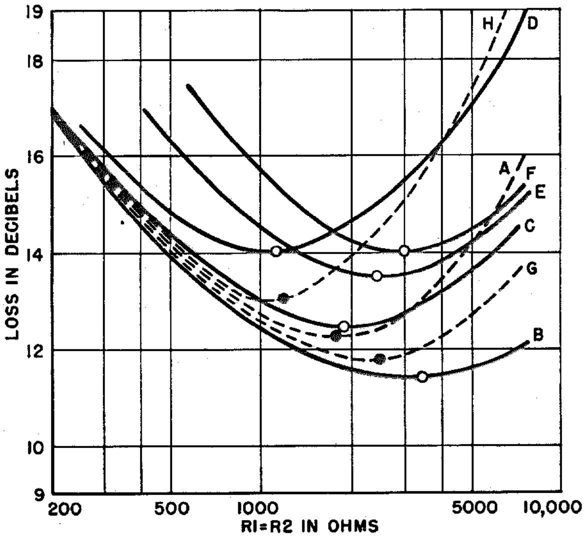


Fig. 10—Compensating circuit for parasitic capacitance.

by Clavier and Denis, has been used. The whole circuit, including the parasitic capacitance, has a parallel resonance at 90 kilocycles and a sufficiently high impedance at 0.1 kilocycle. Its effect on the loss of a modulator working between equal pure resistances has been measured and can be seen by comparing Curves *A* and *B* of Fig. 11. Curve *A* is identical to Curve *A* of Fig. 8, and gives the performance of the original modulator; Curve *B* is the attenuation of the compensated modulator. The accuracy of the compensation has been checked by adding to the compensated circuit an external capacitance of



	A	B	C	D	E	F	G	H
r IN OHMS	0	0	0	0	100	200	0	0
R IN KILOHMS	∞	∞	10	3	10	10	∞	∞
C IN MICROMICROFARADS	1300	0	0	0	0	0	600	2700

Fig. 11—Measured transmission loss of a bridge-type modulator between resistances for various values of r, R, and C (see Fig. 12). The circles and dots indicate the agreement of the calculated values with the measured results.

1300 micromicrofarads. The curve so obtained coincides fairly well with Curve A.

4.3 Verification of Resistive Theory

To check the validity of (7), the forward and reverse resistances have been artificially modified by incorporating in the compensated modulator a series resistance r or a shunt resistance R (Fig. 12), this being equivalent to connecting r or R in series or in shunt with each rectifier.

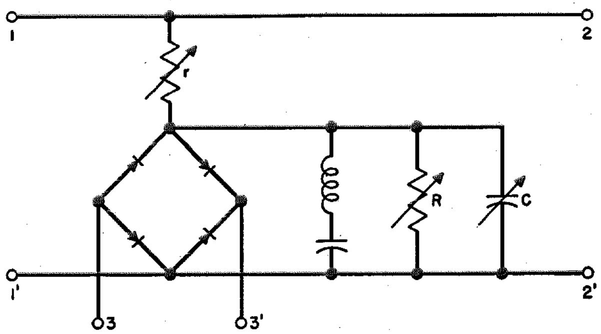


Fig. 12—Measuring circuit for Fig. 11.

A large series of measurements has been carried out with different values of r, R, and with terminating resistances $R_1=R_2$. Some typical curves are plotted on Fig. 11, Curves B, C, D, E, and F. In Table I, the minimum attenuation and the optimum impedance are calculated from (7), using the measured value $r_s=140$ ohms and a value of 18,500 ohms for r_0.

TABLE I

Curve in Fig. 11	r	R	r_s'	r_o'	R_{opt}	$A_{\min}$
B	0	∞	0.14	18.5	3.2	11.5
C	0	10	0.14	6.5	1.9	12.6
D	0	3	0.14	2.6	1.2	14.0
E	0.1	10	0.24	6.5	2.5	13.3
F	0.2	10	0.34	6.5	2.95	13.9

All resistances are in thousands of ohms. $r_s'=r_s+r$, $r_o'=r_oR/(r_o+R)$; $r_s=0.14$, and $r_o=18.5$. $R_{\text{opt}}=2(r_s'r_o')^{1/2}$; $A_{\min}=9.94+17.4(r_s'/r_o')^{1/2}$ decibels.

The calculated results for R_{opt} and $A_{\min}$ agree with the corresponding measured values. To permit easy comparison, they have been denoted by small circles in Fig. 11.

The value of 18,500 ohms for r_o has been adopted as that most in agreement between measured and calculated values. No direct measurement of a mean value of the reverse resistance is possible, but the usual direct-current measurement with a small superimposed alternating voltage yields values of the same order of magnitude.

4.4 Effects of Parasitic Capacitance

Starting with the compensated modulator, the effect of adding a variable shunt capacitance C has been investigated, using the circuit of Fig. 12 with $r=0$ and $R=\infty$. Some of the attenuation curves obtained are shown in Fig. 11, Curves A, G, and H. The supplementary loss A_c, due to the capacitance, is fairly well accounted for by the formula

$$A_c=\tfrac{1}{2}\log\left[1+\left(\frac{RC\Omega}{4}\right)^2\right]\text{ nepers.}\qquad(41)$$

In this case, where the detected frequency is very low compared with the signal and carrier frequencies, or in the opposite case where the signal frequency is very low, such a formula can be roughly established by highly approximate theoretical considerations. The general case is very difficult.

For nearly ideal bridge modulators working between equal pure resistances, $R_1=R_2=R$, and if ω or $\Omega-\omega$ are small compared to Ω, an approximate expression of the total loss is

$$A=\log\pi+\frac{R}{2r_o}+\frac{2r_s}{R}+\frac{1}{2}\left(\frac{RC\Omega}{4}\right)^2\text{ nepers.}\qquad(42)$$

Calculations for the minimum loss lead to a third-degree equation, which can be solved trigonometrically. The optimum value of R is

$$R=\frac{(3r_or_s)^{1/2}}{\cos\frac{\varphi}{3}},$$

$$\cos\varphi=\frac{1}{2}\left(\frac{C\Omega}{2}\right)^2 r_s^{1/2}(3r_o)^{1/2}.\qquad(43)$$

If the right-hand member of the expression of $\cos\varphi$ is larger than 1, the cos must be replaced by cosh in both (43) and (45). For small values of C, (43) gives (5), and for large values of C,

$$R=2r_s^{1/3}\left(\frac{2}{C\Omega}\right)^{2/3}.\qquad(44)$$

The minimum attenuation is

$$A = \log \pi + \frac{1}{4}\left(\frac{3r_s}{r_o}\right)^{1/2} \frac{1+4\cos^2\frac{\varphi}{3}}{\cos\frac{\varphi}{3}}, \qquad (45)$$

which becomes, for large values of C,

$$A = \log \pi + \frac{3}{2}\left(\frac{r_s C\Omega}{2}\right)^{2/3}. \qquad (46)$$

Both (43) and (45) agree with measured results. For an easy comparison, calculated values have been represented by dots on Fig. 11.

5. *Acknowledgment*

The theoretical and practical work described herein was started in 1944 at the suggestion of Mr. L. A. Braem, chief engineer of the transmission department. The author is glad to take this opportunity of expressing his appreciation to Mr. Braem and to several colleagues in the transmission department; especially to Mr. E. Van Derstappen and Mr. Karlin. Unpublished memoranda by Messrs. Clavier and Denis, and by Katchatouroff and Lalande, all of Laboratoire Central de Télécommunications, and discussions with other engineers of that organization, have also been most helpful.

Addendum, Vol. 24, No. 1, 1947

FRANCE–ENGLAND SUBMARINE CABLE (1939) AND PARIS–CALAIS CABLE

On page 26, the following statement is made:

> This is the first cable in which the two transmission directions of one 12-channel link have been included under a common lead sheath; and, so far as is known, it is the only cable of its kind in Europe or in America.

Bell Telephone Laboratories have kindly informed us that:

> A layer-shielded cable for opposite-directional 12-channel carrier operation was installed in Oregon in 1938. The length was about 51 kilometers.

No information on the structure, characteristics, or methods used to compensate for cross talk has been published on this cable.

Accuracy of Impedance Measurements

By B. SECKER

Standard Telephones and Cables, Limited, London, England

MEASUREMENTS of the impedance of networks are frequently made in the telecommunications art. Attention is drawn to a number of important factors that affect the successful use of impedance bridges. For a large number of impedance measurements, it is unnecessary to know the exact value of the impedance, but rather the degree of unbalance that exists between the impedance under test and some other impedance. Impedance measurements may, therefore, be divided into two categories; impedance-unbalance measurements and true impedance measurements. Sources of errors in bridges of various types are discussed, and comparisons made between different bridges for particular measurements. Curves show the limits of measurement for series- and parallel-resonance bridges, and the fundamental balance conditions for different types of bridges are derived.

• • •

It frequently happens that the user of an impedance bridge is comparatively unfamiliar with the principles governing the operation of the bridge he is using, and the variety of circuits employed in impedance bridges is so extensive that a complete education of the user in bridge technique would be a process both tedious and uneconomical. It is disappointing and exasperating to the user to be informed, after long and laborious measurements, that the bridge he has been using either does not cover the range of measurement required, or is otherwise liable to give inaccurate results.

Individual bridges have been developed for special purposes, and contain compensating circuits enabling many sources of errors to be ignored over a restricted range of measurement. Outside this range, however, such compensating circuits often increase the inaccuracy of the bridge. It is therefore important, before beginning to make measurements, to consider carefully the nature of the impedance under investigation, and the characteristics of the bridge circuit proposed for use.

1. Impedance

Strictly speaking, impedance is an extension of the resistance concept embodied in Ohm's law. But, because varying (sinusoidal) currents are involved, the impedance of a network is affected by its surroundings to a great extent, and it is important that these effects should be either eliminated or estimated, if accuracy of measurement is required.

To get a clear idea of the method of measurement to be employed, it is essential to consider the reason for making the measurement. In many cases, the significant information required is the unbalance between two impedances. This is the case where the termination of a line is under consideration. A standard impedance is often selected, and all other networks are adjusted against the standard to keep the impedance unbalance within certain limits. Such a requirement is really an "acceptance" test and, if the impedance unbalance can be measured directly, it is more economical to do so than to make impedance measurements. Direct measurement, however, may give rise to a number of errors which will be discussed later.

On the other hand, the measurement of impedance unbalance against a known impedance gives very little information on the network as an individual piece of apparatus. Precise impedance measurement, however, will indicate a line of action when investigating faults within a network. With negative-feedback amplifiers, for example, the effect of variations in the feedback circuit on the input impedance of the amplifier is of importance. Many other cases of this sort could be cited in which the essential feature is the estimation of the degree of dependence of an impedance upon some other factor. The mode in which the impedance varies can then be used to provide information in the search for faulty circuit components.

It is, therefore, clear that impedance measurements can be divided roughly into two groups: "acceptance" tests, and precise measurements. These will be considered in turn in Sections 2 and 3.

2. *Impedance Unbalance*

The quantity known variously as impedance unbalance, singing point, and return loss, together with its associated function, the reflection coefficient, can be measured directly. The usual way is to use an impedance bridge with one impedance as standard. Instead of balancing the bridge, the unbalance voltage at the detector is compared with that due to a pair of impedances of known unbalance.

Most bridge errors result from the various constants of the bridge, and a null method normally eliminates them when an initial zero balance is made. In the unbalance method, this is not necessarily so, and the major source of error is this lack of compensation. To overcome this difficulty, it is usual to define the calibrating unbalance at the "acceptance" point. In other words, if 26 decibels is the acceptance limit of unbalance, two resistors having that unbalance are used for calibration. They are, furthermore, selected to have approximately the same impedance as the network under test. In this way, the effect of strays is eliminated at the critical point. At other points, the errors are not important.

However, such apparatus is highly specialised. If there is any doubt as to the suitability of enclosed standards of calibration, it is much better to make up a pair of external standards. In this way, inaccuracy at the critical point is almost eliminated.

The calibrating resistances normally incorporated in impedance-unbalance bridges are such as to produce an unbalance of 25 decibels for apparatus designed to measure reflection factor in 600- and 125-ohm circuits over the frequency range of 1 to 150 kilocycles. For apparatus designed to measure 600-ohm circuits in the frequency range from 50 cycles to 15 kilocycles, the calibrating resistances are usually selected so as to produce an unbalance of 30 decibels. It is, therefore, important before using any impedance-unbalance bridge, first to ascertain the impedance for which the bridge has been designed and the degree of unbalance used to calibrate it, and then to determine the effects that may be produced by the circuit under investigation, and by the oscillator and detector used for the measurement.

Formulas covering the oscillator and detector impedances, and the coupling coefficients between the various windings of the hybrid coil, where this method of measurement is used, have been published.[1] If a plain Wheatstone's bridge is used for the measurement, appropriate formulas will be found in Appendix *E*, Section IV of reference 1.

3. *Impedance Measurement*

Broadly speaking, impedance measurements by null methods fall into two sub-groups, one being concerned with the measurement of almost pure reactances, while the other deals with comparatively wide variations of phase angle.

The term *phase angle* is a deplorable one for the angle whose tangent is the ratio of reactance to resistance, but in default of an equally short term for this quantity, it is proposed to make use of it. In view of the fact that questions affecting wave transmission have no part in this paper, no confusion of meaning should arise.

When dealing with almost pure reactances, it is usual to balance or neutralise the large component by an accurately calibrated standard, and then to measure the small component as a secondary effect. In this way, the errors due to impurity in the secondary standard have third-order effects on the large component.

3.1 INDUCTANCE MEASUREMENT

As a first example, the measurement of inductance may be considered. In a series-resonance bridge, a good air capacitor will have a small effect on the resistance of the coil under test. The effective resistance will be slightly increased by it, and the total resistance can be measured by a resistance box in the other arm of the bridge. The self-inductance and capacitance of this box will be small when compared with the reactance of the coil. In this way, the error in inductance measurement will be second order, and negligible.

Section 7.1 shows that, if an inductance is measured by means of a series-resonance bridge, $Q\ (=L\omega/R)$ should be increased by a factor

$$\frac{\theta}{Q}+\phi Q,$$

---

[1] K. S. Johnson, "Transmission Circuits for Telephone Communication," D. Van Nostrand, New York, 1929; Appendix *E*, Section IX.

in which θ is the time constant of the resistance standard, and ϕ the power factor of the capacitance standard.

The series-resonance bridge, therefore, gives a value for Q that is lower than the actual value.

The inductance of the coil being measured is given in Section 7.1 as

$$L=\frac{1}{C\omega^2}(1+\theta R_oC\omega);$$

immediately following this it is shown that, to a first approximation,

$$Q=\frac{1}{R_oC\omega},$$

so that the factor by which the bridge readings must be multiplied is

$$1+\theta R_oC\omega=1+\frac{\theta}{Q}.$$

Formulas are derived in Section 7.2 for the resistance and inductance of the same coil measured in a Maxwell-type bridge using similar standards. The magnitude of the errors is the same, but the correction due to the power factor of the capacitor is negative.

This bridge, therefore, gives apparent values of Q that are higher than the true values. The correction factor is

$$1-\frac{\theta}{Q}-\phi Q.$$

Table I gives the comparative errors introduced by the two bridges in the measurement of a coil for which $Q=100$. It is assumed that $\phi=\theta=0.001$.

TABLE I

COMPARATIVE ERRORS IN INDUCTANCE MEASUREMENT

Quantity	Correction Factor	
	Maxwell Bridge	Series-Resonance Bridge
L	−0.00001	+0.00001
R	+0.1	−0.1
Q	−0.1	+0.1

Parallel-resonance bridges give similar errors to those given by the series-resonance type, i.e., the measured Q of a coil is too low.

From the above discussion, it can be seen that even if the bridge network has no strays, has perfect ratio (or product) arms, and ideal shielding, it is possible to get considerable variations in measurements with different bridge circuits as a result of comparatively slight imperfections in standards.

3.2 CAPACITANCE MEASUREMENT

In general, two circuits are used for capacitance measurements. Fig. 3* illustrates both. R and R are ratio arms, one of which has a capacitance C_1 in parallel. The standards consist of a capacitance C and a conductance G in parallel. The inductance and resistance of leads are represented by L and r, respectively, and ϕ is the imperfection of the standard capacitor.

For a capacitance-and-conductance bridge, these quantities are differential, the standards being changed from one arm to the other to obtain balance. C_1 is assumed to be zero.

In the Schering circuit, C_1 is used to measure power factor and G is assumed to be zero. The initial capacitance between the test terminals is kept as low as possible by careful shielding.

Comparing these two circuits, it can be seen that the Schering bridge is quite unsuitable for measuring capacitances that are balanced to earth, as this entails substantial admittances to earth from the test terminals. On the other hand, the capacitance-and-conductance bridge has no such restrictions.

However, the formulas for the accuracy of the two bridges given in Section 7.3 have a number of interesting features.

First, in both bridges there is a multiplying factor for capacitance. This can be eliminated by calibrating the bridge against external standards.

Secondly, the Schering circuit requires a correction factor $C/(C-C_o)$, which is usually nearly unity.

Finally, the stray inductances, resistances, and imperfection in the standard capacitor give additive errors. These are $\phi+r\omega(C+C_o)$ for the capacitance-and-conductance bridge, and $rC\omega$ for the Schering bridge; ϕ is the weighted power factor of the standard capacitor, and C_o is the initial capacitance.

The factor ϕ also appears in the exact formula for the Schering, but only as the difference between the power factors at the two settings of the standard.

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*All figures will be found in Section 7, Appendix.

It should be noted, in any case, that the capacitance-and-conductance bridge has an error $\omega r C_o$ in excess of the Schering bridge, and since C_o is quite large in the capacitance-and-conductance bridge, this error is considerable.

It would appear that for measuring two-terminal capacitors, the Schering bridge has distinct advantages. It does not have the versatility of the capacitance-and-conductance bridge for "balanced-to-earth" measurements.

3.3 Impedance Measurement

Finally, the measurement of general impedance may be considered. In the cases previously discussed, the major difficulty has been to obtain the value of a small component of an impedance with any reasonable accuracy. The problem which now presents itself is the measurement of impedance of any magnitude, and the fact that reactance is dependent on frequency imposes a definite limitation on the range of measurement. However, most impedance measurements require vectorial accuracy rather than the accurate measurement of small components, and this fact provides some mitigation of the difficulties associated with the design of impedance bridges.

The selection of standards for the bridge is the most important consideration. Resistance standards for the measurement of real components introduce difficulties as a result of their frequency characteristics. The pure resistance is an unattainable ideal, particularly at the higher frequencies. The standards used for the measurement of reactive components are normally inductances or capacitances. While it is possible to provide fixed mutual-inductance standards with constants calculated with great accuracy, and suitable for measurements over a restricted frequency range, such standards belong to the domain of primary standards, and their cost prohibits their employment in commercial types of impedance bridges. Recourse is, therefore, generally made to capacitance standards, and it is fortunate that first-class capacitance standards are easy to obtain. These standards are also stable, they hold their calibration, and the imperfections in the smaller values can be made so small that there is now a tendency to use circuits that employ capacitors for the measurement of resistance values.

The normal circuits used for impedance measurement fall into two main divisions: series capacitance-and-resistance combinations, and parallel arrangements. Further difficulties are now encountered. The reactance of a capacitor cannot vary continuously from zero, nor can the conductance of a resistance box. This means that a series combination must start from a false reactance zero, and a shunt arrangement from a false conductance zero, if an effective zero balance is to be obtained.

The usual way to ensure this is to include a fixed capacitor permanently in series with the unknown for series measurements, and a fixed resistor across the unknown terminals in shunt arrangements. The latter can be disconnected when measuring, if desired.

The series arrangement has the advantage that the results obtained are in the conventional form $R+jX$. The chief disadvantages are that the circuit can be adapted to "balanced-to-earth" measurements only with great difficulty, and that there is a lack of symmetry about the zero of the reactance scale. The curve in Fig. 5 shows this effect. It is obvious that a linear scale in capacitance makes for easy adjustment at the extreme positive-reactance end, while at the extreme negative portion, adjustment will be almost impossible. In such bridges, therefore, there will be an apparent loss of sensitivity as the reactance to be measured passes from a negative value through zero to positive. If only a single air capacitor is used, this effect can be partially eliminated by appropriate shaping of the plates. If the range of measurement of such a bridge is plotted as an area on a plane, the resistive component being horizontal and the reactive vertical, a rectangle is obtained, Figs. 7 to 11, which is bounded on the right by the maximum value of the standard resistance. The horizontal boundaries of the rectangle will depend on the curve shown in Fig. 5, and will vary with frequency. Figs. 7 to 11 show the bridge limits at frequencies of 50, 80, 500, 1000, and 3000 kilocycles.

In the case of the parallel arrangement, on the other hand, capacitance is effectively an admittance, and there will be no skewness in the capacitance calibration when passing from positive to negative angles. However, the range of impedance covered has a rather complicated boundary. A circle described on the lowest value of resistance in the conductance box is the inner limit. The other portion of the boundary is a pair of semicircles with points on the imaginary axis as centres. These form a series of three arcs, at the left of which no balance can be obtained. The semicircles are dependent on frequency and, therefore, complicate matters when considering the range of measurement available.

Figs. 7 to 11 illustrate these facts in a striking manner. They each show at one frequency the respective measuring ranges of two complementary bridges. One of these is a series bridge consisting of a resistance decade of 0 to 100 ohms in series with a capacitor having a range of 1000 to 20,000 micromicrofarads. The offsetting capacitance is 5000 micromicrofarads. The other bridge has a conductance of 0 to 11,000 micromhos shunted by ±11,000 micromicrofarads. A casual examination of these bridges gives the impression that practically all of the impedance field must be covered, but in Figs. 7, 10, and 11 the considerable ranges shown shaded are not covered by either bridge at 50, 1000, and 3000 kilocycles. Whatever the quality of the components used on this pair of bridges, the effective frequency range is reduced to 80 to 500 kilocycles, and it is difficult to see how to extend the range of such combinations of bridges.

With regard to inaccuracy, it is enough to observe that the chief causes are impurities in the resistance and capacitance standards. Section 7.4 shows them as follows:

For impedances, $Z=R+jX$.
For admittances, $Y=G+jB$.

The impurity in capacitance is $j\phi C$, and in resistance is $j\theta R$. Then the true values are given by

$$Z=R\left(1-\phi\frac{X}{R}\right)+jX\left(1+\theta\frac{R}{X}\right)$$

and

$$Y=G\left(1-\phi\frac{R}{G}\right)+jB\left(1-\theta\frac{G}{B}\right).$$

4. *Errors Due to Leads*

In the preceding section, the effect of inductance and resistance in the leads to the impedance under test has been ignored. The reason for this is quite simple. Reactances can usually be brought to a bridge, but impedances often have to be measured *in situ*. This means bringing the bridge to the unknown, and involves considerable modifications in bridge design. Usually, a test head is used with a long lead back to the bridge. A similar lead can then be included from bridge to standards. Such a device requires that the test head shall transform the impedance to be measured into one whose form can be simulated by the standards. It is useless to use a parallel-resonance method, for example, with the capacitance and inductance separated by a long lead. In these circumstances it is usual to include offsetting capacitors in the test head.

Formulas can be derived for the measurement of impedance through a long lead, or through any other four-terminal network, but they are not popular with users. In general, it is preferable to use compensating leads and a test head.

5. *Miscellaneous*

So far, attention has been confined to the consideration of errors and to the difficulties arising from the standards and the leads to them. Other factors may now be considered.

First, ratio and product arms. Errors in these can cause considerable trouble, especially when reactances are considered. A difference in phase angle introduces a large error in the small component. For example, in a Maxwell bridge at a frequency of 750 kilocycles, a ratio arm of 333 ohms, which has an uncompensated capacitance of as little as 3 micromicrofarads, can change the apparent Q of a coil from a true value of 200 either to 100 or to infinity, according to the direction in which the capacitance affects it.

A further point is the question of "balanced-to-earth" measurements. In such measurements, the actual capacitance measured depends on the admittances of the bridge points to earth. If these are large compared with those of the network under test, an "enforced centre point" is obtained. For example, a capacitance C with C_1 and C_2 as capacitances to earth from its terminals, will have some apparent capacitance ranging from

$$C+\frac{C_1+C_2}{4}$$

with large bridge admittance, to

$$C+\frac{C_1C_2}{C_1+C_2}$$

with zero bridge admittance. These are equal when the test impedance is perfectly balanced to earth; when $C_1 = C_2$.

The advantages of using inductive ratio arms are not generally appreciated. Inductive ratio arms permit of a higher degree of balance than that obtainable with resistance arms. Furthermore, by connecting the oscillator to the junction point of the two ratio arms (i.e., between points B and D in Fig. 6, the potentials at points A and C are practically equal to that at B due to the fact that the windings of the two ratio coils are mutually opposing. This also implies an absence of flux in the core, and therefore no danger of distortion to the oscillator wave form. The potentials at A and C must be equal at the balance point, consequently practically the entire oscillator voltage appears across the unknown, making for great sensitivity in the neighbourhood of the balance point.

If point B is earthed and the unwanted terminals of the network under test connected to it, any capacitance between point C of the unknown and earth is ineffective, since there is no potential across it, while capacitances between point D and earth are effectively connected across the oscillator, and are equally harmless. Only a single reading is therefore necessary, instead of two, for the determination of direct capacitance.

Shielding is also very important for eliminating induction and strays. Earthing paths can give a great deal of trouble. Wherever possible, it is best to earth everything, oscillator, detector, etc., via the bridge network.

6. *Conclusion*

It is at once obvious from the foregoing discussion that accuracy can be obtained only if great care is taken. It has been seen that correction factors can have considerable effect on results, and it follows from this that direct-reading bridges, suitable for semiskilled operation, will require considerable compensation. If such apparatus is to be reliable, a routine of regular adjustment must be set up, such adjustments being carried out by skilled personnel.

When a high degree of accuracy is required, the less permanent the compensation, the better the results. Direct adjustment of the compensating circuits by the user is necessary. The measurement of leads and other relevant factors will have to be made individually, and the correction factors calculated. It is sometimes of advantage to reverse the ratio arms, if possible. The effect of direct induction between oscillator and detector should be assessed by reversing the leads between the oscillator and the bridge, and between the bridge and the detector. The effect of varying the method of earth connection, and of changing amplitude of the oscillator input to determine nonlinear effects in the test impedance, should also be studied.

7. *Appendix*

7.1 SERIES-RESONANCE BRIDGE

In the series-resonance bridge, shown in Fig. 1, equal inductive ratio arms are used.

The imperfect standard resistance is denoted by

$$R_o(1+j\theta),$$

and the imperfect capacitance standard is denoted by

$$C(1-j\phi).$$

The unknown is an inductance L with effective resistance R.

The balance condition is approximately

$$R+jL\omega+\frac{1}{jC\omega}(1+j\phi)=R_o(1+j\theta),$$

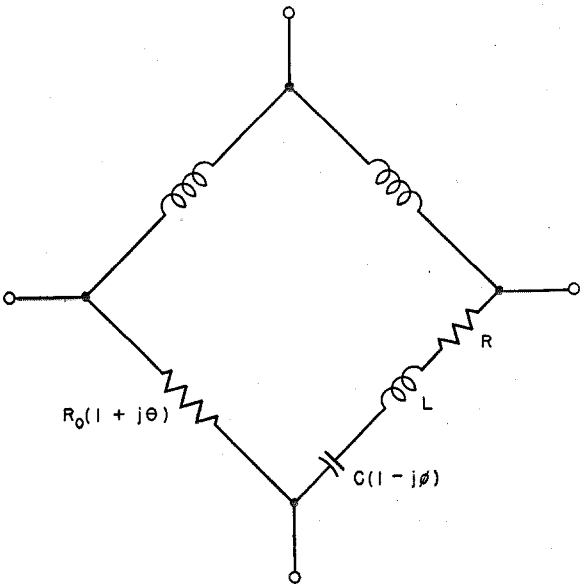


Fig. 1—Series-resonance bridge.

or

$$R=R_o-\frac{\phi}{C\omega}$$

and

$$LC\omega^2=1+\theta R_oC\omega.$$

$$Q=\frac{L\omega}{R}=\frac{1}{R_oC\omega}\left(\frac{1+\theta R_oC\omega}{1-\dfrac{\phi}{R_oC\omega}}\right)=\frac{1}{R_oC\omega}\left(1+\frac{\theta}{Q}+\phi Q\right)\cdot$$

7.2 MAXWELL BRIDGE

In the Maxwell bridge, shown in Fig. 2, similar standards to those of Section 7.1 above are used.

Then,

$$R+Lj\omega=G(1-j\theta)+jC\omega(1-j\phi),$$

in which R_o is written as a conductance G. This gives

$$R=G(1+\phi Q),$$

$$L=C\left(1-\frac{\theta}{Q}\right),$$

and

$$Q=\frac{L\omega}{R}$$

$$=\frac{C\omega}{G}\left(1-\frac{\theta}{Q}-\phi Q\right)\cdot$$

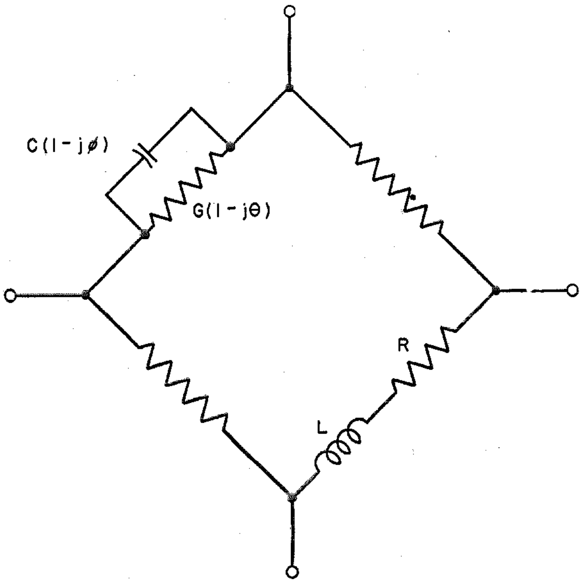


Fig. 2—Maxwell bridge.

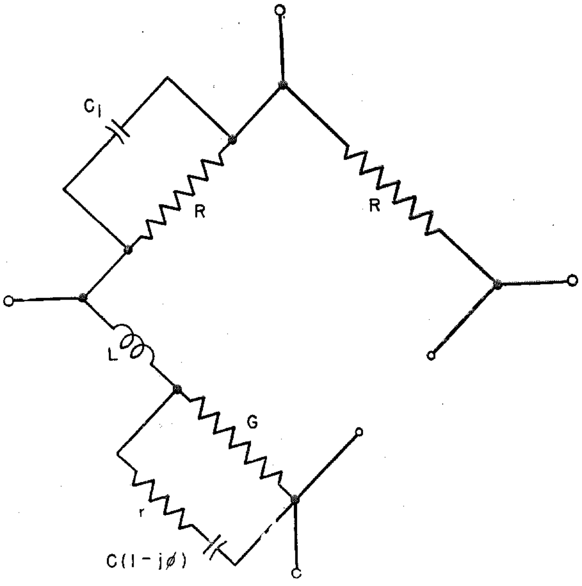


Fig. 3—Capacitance bridge.

7.3 CAPACITANCE BRIDGE

Fig. 3 shows a general form of the capacitance bridge. It can be used as a Schering or as a capacitance-and-conductance bridge by changing the method of balancing the conductance.

To the first order of approximation,

$$C_x=C(1+LC\omega^2),$$

and

$$G_x=G+\phi C\omega+rC^2\omega^2+RC_1C\omega^2.$$

The difference in the two types arises from the methods of conductance and zero balance employed.

7.3.1 Capacitance-and-Conductance Bridge

Initial conditions:

Capacitance $=C_o(1+LC_o\omega^2)$.

Conductance $=G_o+\phi_oC_o\omega+rC_o^2\omega^2+RC_1C_o\omega^2$.

Therefore,

$$C_x=(C-C_o)[1+L\omega^2(C+C_o)],$$

and

$$G_x=G-G_o+\omega(\phi C-\phi_oC_o)+r\omega^2(C^2-C_o^2)$$
$$+RC_1\omega^2(C-C_o).$$

Assuming that $C_1=0$ and $\phi=\phi_o$,

$$C_x=(C-C_o)[1+L\omega^2(C-C_o)],$$

$$G_x=G-G_o+\omega(C-C_o)\phi+r\omega^2(C^2-C_o^2),$$

and

Power Factor

$$=\frac{1}{1+L\omega^2(C_o+C)}\left[\frac{G-G_o}{(C-C_o)\omega}+\phi+r\omega(C+C_o)\right].$$

7.3.2 *Schering Bridge*

Initial conditions:

$$G=G_o=0.$$

$$\text{Capacitance}=C_o(1+LC_o\omega^2).$$

$$\text{Conductance}=\phi_oC_o\omega+rC_o^2\omega^2+RC_{10}C_o\omega^2=0.$$

Then,

$$C_x=(C-C_o)[1+L\omega^2(C+C_o)],$$

and

$$G_x=C\omega[\phi-\phi_o+r\omega(C-C_o)+\omega R(C_1-C_{10})].$$

If $\phi=\phi_o$,

Power Factor

$$=\frac{1}{1+L\omega^2(C_o+C)}\left[rC\omega+\frac{RC\omega(C_1-C_{10})}{C-C_o}\right].$$

The normal correction factor is

$$\frac{C}{C-C_o}.$$

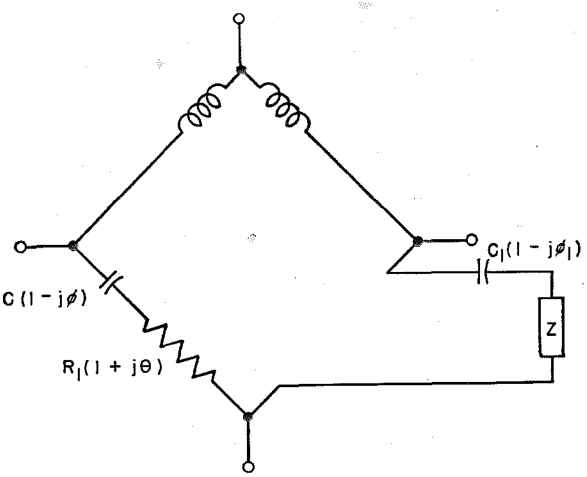


Fig. 4—Series-impedance bridge.

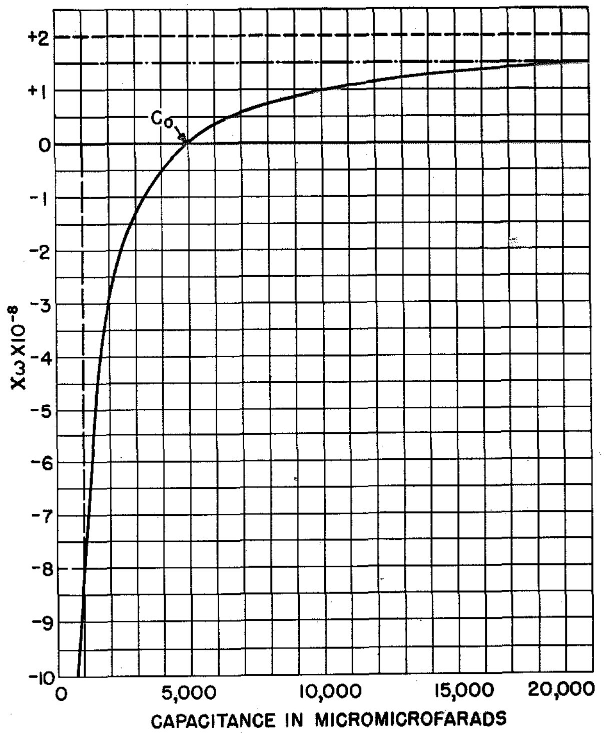


Fig. 5—Measurement of impedance with a series-impedance bridge, using 1000-to-20,000-micromicrofarad standard capacitor.

- - - - - positive limit of X.
—·—· upper limit of measurable range.
— — — lower limit of measurable range.

7.4 SERIES-IMPEDANCE BRIDGE

With the offsetting capacitor as shown in Fig. 4,

$$Z=R+jX$$

$$=R_1(1+j\theta)+\frac{1}{jC\omega(1-j\phi)}-\frac{1}{jC_1(1-j\phi_1)\omega},$$

$$R=R_1+\frac{1}{\omega}\left(\frac{\phi}{C}-\frac{\phi_1}{C_1}\right),$$

and

$$X=\frac{C-C_1}{CC_1\omega}+R_1\theta.$$

If an initial reading is taken with Z short circuited,

$$0=R_{10}+\frac{1}{\omega}\left(\frac{\phi_o}{C_o}-\frac{\phi_1}{C_1}\right)$$

$$=\frac{C_o-C_1}{C_oC_1\omega}+R_{10}\theta_o.$$

If $\theta=\theta_o$, and $\phi=\phi_o$, and all are small,

$$R=(R_1-R_{10})-\phi X,$$

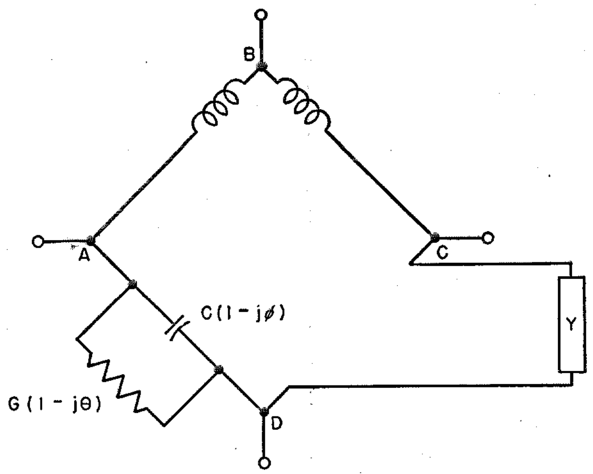


Fig. 6—Parallel-admittance bridge.

and

$$X = \frac{C - C_o}{CC_o\omega} + R\theta.$$

The curve $X\omega = \dfrac{C - C_o}{CC_o}$ is shown in Fig. 5, and indicates clearly the apparent loss of sensitivity at the positive end of the scale.

7.5 Parallel-Admittance Bridge

This bridge is illustrated in Fig. 6; the zero adjustments are straightforward, so

$$\begin{aligned} Y_x &= G_x + jB_x \\ &= G(1 - j\theta) + j\omega C(1 - j\phi), \end{aligned}$$

from which

$$G_x = G + \phi\omega C,$$

and

$$B_x = \omega C - \theta G.$$

It is obvious from these two sets of equations that the errors due to impurity in standards modify results as follows:

$$Z = R\left(1 - \phi\frac{X}{R}\right) + jX\left(1 + \theta\frac{R}{X}\right),$$

and

$$Y = G\left(1 - \phi\frac{B}{G}\right) + jB\left(1 + \theta\frac{G}{B}\right)\cdot$$

The vector field covered by a typical pair of bridges is shown in Figs. 7 to 11 for frequencies between 50 kilocycles and 3 megacycles.

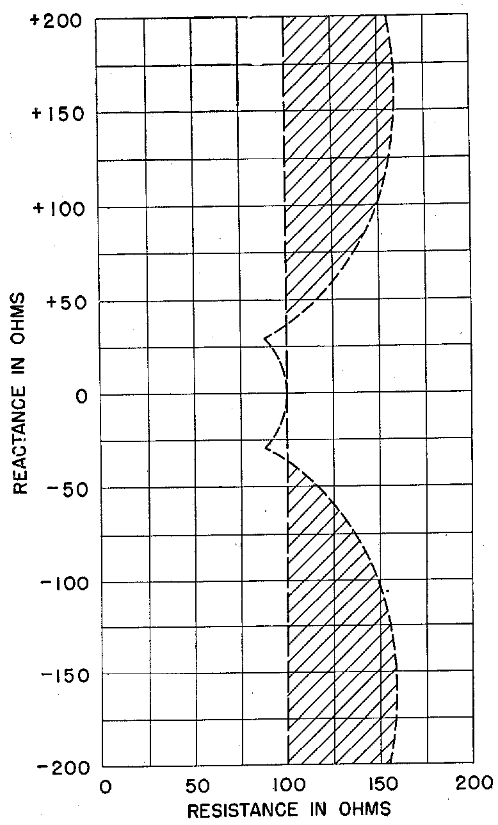


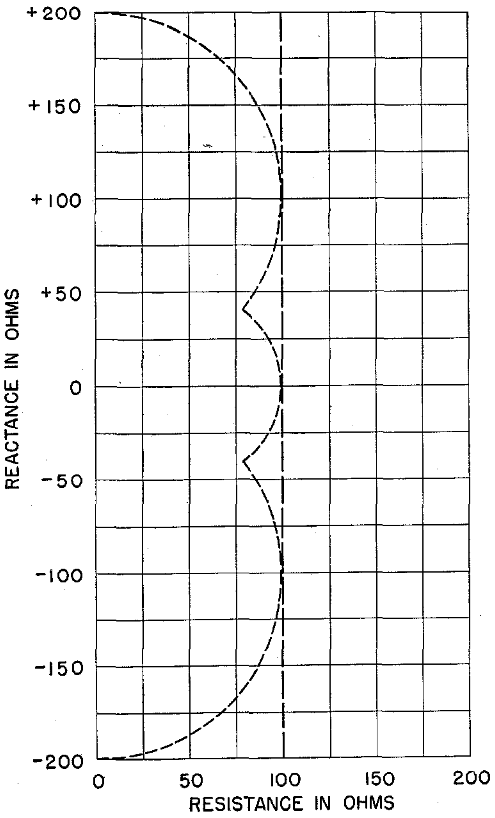
Fig. 7—Unmeasurable range of impedances, not covered by use of either a series-impedance or parallel-admittance bridge at 50 kilocycles, is indicated by the shaded area. The line of long dashes indicates the upper limit of the series bridge, and the line of short dashes the lower limit of the parallel bridge.

Fig. 8—Same as Fig. 7, but measured at 80 kilocycles. Shown on opposite page upper left.

Fig. 9—Same as Fig. 7, but measured at 500 kilocycles. Shown on opposite page lower left.

Fig. 10—Same as Fig. 7, but measured at 1 megacycle. Shown on opposite page upper right.

Fig. 11—Same as Fig. 7, but measured at 3 megacycles. Shown on opposite page lower right.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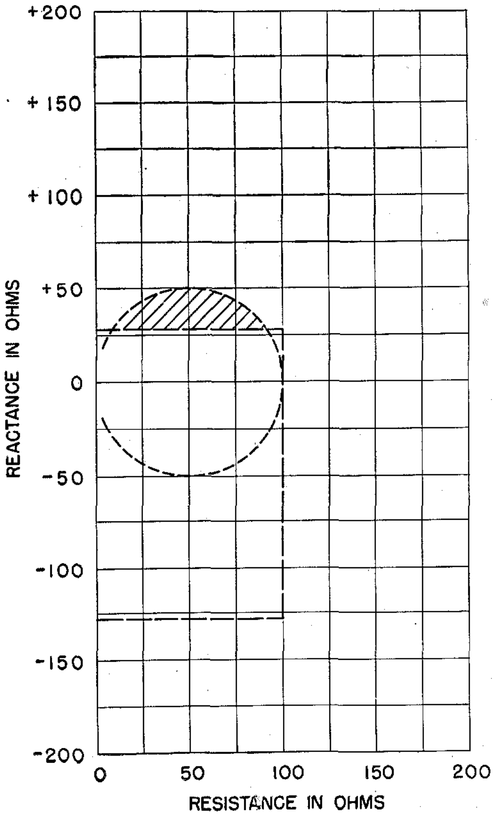
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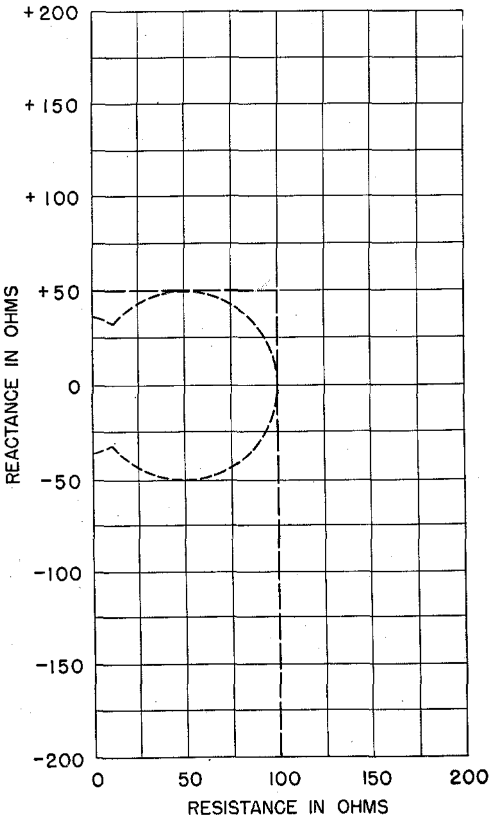
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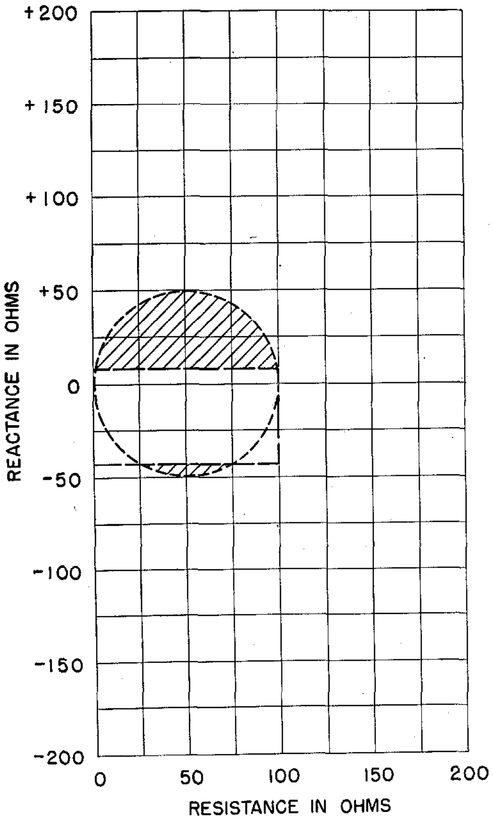
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Heating of Radio-Frequency Cables

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HEATING of radio-frequency cables is developed in this paper on basic physical phenomena. Only the simplest physical or mathematical assumptions are made, and the mathematical details are worked out with the physical picture in mind. Consequently, it will seem to some readers that many of the steps could have been bridged by the direct application of results of the usual methods of analysis. However, this would make the paper less intelligible to others, and would dull the over-all physical picture. The general treatment of such problems has been developed by J. B. J. Fourier[1] (1768–1830), and by numerous others in works on partial-differential equations as applied to such subjects as heat, sound, and electricity.

• • •

The power-handling capacity of a radio-frequency cable or transmission line is determined by the maximum voltage that the dielectric (whether air or solid) can withstand continuously without breakdown, and the maximum temperature rise that the components of the cable can withstand without chemical deterioration or significant physical change. The problem of heat in radio-frequency cables is discussed here with particular reference to the solid-dielectric coaxial type. The problem resolves itself into two parts; the electrical one in which the generation of heat is determined with respect to location and intensity, and the thermal one concerning the resulting temperature rise at each point in the cable and the dissipation of heat. The heating of the cable by external sources is also of importance.

In general, investigation of the steady-state conditions is sufficient, as the maximum temperature rise is of principal importance. However, in the case of pulsed transmission or keying, the steady state is composed of a succession of transients, all such transients usually being assumed to be identical in character and repeated at a constant rate. The temperature rise then of importance is the highest during any one pulse.

1. Symbols

The symbols listed below appear more or less generally throughout the paper. Others used briefly are defined where introduced, if their identities are not obvious. Unit length may be assumed to be 1 centimeter, and unit temperature difference, 1 degree centigrade. Subscripts 1 and 2 refer to the inner and outer conductors, respectively, while subscript p refers to the dielectric. All logarithms are to the base $e=2.718$.

1.1 MECHANICAL SYMBOLS

A_1 = cross-sectional area of inner conductor.
A_2 = cross-sectional area of outer conductor.
m_c = mass of conductor per unit volume (density).
m_p = mass of dielectric per unit volume (density).
r = radius of a point in the dielectric core.
r_1 = outside radius of inner conductor.
r_2 = inside radius of outer conductor.
r_3 = inside radius of jacket.
r_s = radius of outside surface of the cable.
x = distance along cable, measured toward the load from a voltage maximum of the standing waves (midway between adjacent voltage minimums).

1.2 ELECTRICAL SYMBOLS

C = capacitance in farads of unit length of cable.
ϵ = dielectric constant of cable core.
E = root-mean-square voltage at any location in cable.
E_a = root-mean-square voltage of forward-wave train.
E_b = root-mean-square voltage of back- or reflected-wave train.
E_{cr} = root-mean-square voltage at crest of standing wave.
E_{tr} = root-mean-square voltage at trough of standing wave.

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[1] J. B. J. Fourier, "Analytical Theory of Heat," 1822. English translation by A. Freeman, Cambridge University Press, Cambridge, England; 1872.

E_{flat} = root-mean-square voltage when standing-wave ratio = 1.0.

e = instantaneous voltage (subscripts same as for E above).

f = frequency in cycles per second.

i = instantaneous value of current (subscripts as for E).

I = root-mean-square current at any location in cable (subscripts as for E).

I_ϵ = root-mean-square current flowing radially in dielectric core.

λ = electrical wavelength of cable in centimeters $= 3\times10^{10}/f\sqrt{\epsilon}$.

(pf) = power factor of the dielectric.

$\psi = 4\pi x/\lambda$.

R_1 = resistance per unit length of inner conductor at frequency f.

R_2 = resistance per unit length of outer conductor at frequency f.

(swr) = standing-wave ratio $= E_{cr}/E_{tr}$.

$\omega = 2\pi f$.

W = net power delivered by cable to load.

Z_0 = surge impedance of cable.

1.3 THERMAL SYMBOLS

$\mathcal{E}$ = effective total emissivity (radiation and convection) of outer conductor acting through the jacket; in calories per second per unit area of outside surface of cable for unit temperature difference of outer conductor above ambient.

$\mathcal{E}_s$ = total emissivity of outside surface of cable; in calories per second per unit area of surface for unit temperature difference of the surface above ambient.

k_c = heat conductivity of conductor; in calories per second flow across unit cross-sectional area for unit temperature gradient normal to the cross-section.

k_p = heat conductivity of dielectric core material; in same units as k_c.

k_s = heat conductivity of jacket material; in same units as k_c.

q = rate of heat generation or flow; in calories per second.

$q_1 = q_1' - q_1'' \cos\psi$ = rate of heat generation on inner conductor; in calories per second per unit length.

$q_2 = q_2' - q_2'' \cos\psi$ = rate of heat generation on outer conductor; in calories per second per unit length.

$q_p = q_p' + q_p'' \cos\psi$ = rate of heat generation in a section of dielectric; in calories per second per unit length of section.

s_c = specific heat of conductor material; in calories per gram per degree temperature rise.

s_p = specific heat of dielectric core material; in same units as s_c.

u = temperature at any point referred to an arbitrary zero, which is usually the ambient unless otherwise indicated in the context. (For subscripts see paragraph below.)

u_0 = ambient temperature.

U = maximum value of temperature rise when $u = U\cos\psi$. (For subscripts see paragraph below.)

${}_nU$ = coefficient of the nth term of a series.

In order to specify the location of a temperature, or temperature rise u, and the heat source that causes the rise, the following system of subscripts and primes is used: The first subscript numeral or letter refers to the location of the temperature rise, while the second refers to the source of heat generation causing the rise. When a single prime is used, the rise is due to the part of the heat generation which is uniform over the length of cable in the vicinity of the point in question. A double prime refers to the rise due to the part of the heat generation which is a function of x or ψ. Thus,

u_{12} = rise of inner conductor due to heat generated on outer conductor.

u_{21}' = rise of outer conductor due to q_1'.

u_{p1}'' = rise of dielectric due to q_1''.

U_{11} = maximum temperature of inner conductor due to q_1''.

2. *Electrical Problem*

Although the losses in a cable cause both attenuation and the generation of heat, in most practical cases the attenuation per half wavelength is so small that for present purposes it can be neglected in writing the voltage and current relations of the transmission line. The line then carries a "forward" wave traveling from the source toward the load,

$$e_a=\sqrt{2}E_a\cos\left(\omega t-2\pi\frac{x}{\lambda}\right), \tag{1A}$$

$$i_a=\sqrt{2}I_a\cos\left(\omega t-2\pi\frac{x}{\lambda}\right), \tag{1B}$$

and a "back" or reflected wave traveling from the load toward the source,

$$e_b=\sqrt{2}E_b\cos\left(\omega t+2\pi\frac{x}{\lambda}\right), \tag{1C}$$

$$i_b=-\sqrt{2}I_b\cos\left(\omega t+2\pi\frac{x}{\lambda}\right), \tag{1D}$$

where

$$\left.\begin{array}{l} I_a=E_a/Z_0, \\ I_b=E_b/Z_0, \end{array}\right\} \tag{1E}$$

and x is the distance along the transmission line measured from a voltage maximum to the point in question, x being positive toward the load.

These voltages and currents and their resultants are shown in vector form in Fig. 1. The root-mean-square values of the forward and back voltages E_a and E_b are related to the crest and trough voltages of the standing waves along the line by

$$E_{\mathrm{cr}}=E_a+E_b, \tag{2A}$$

$$E_{\mathrm{tr}}=E_a-E_b, \tag{2B}$$

$$E_a=\tfrac{1}{2}(E_{\mathrm{cr}}+E_{\mathrm{tr}}), \tag{2C}$$

$$E_b=\tfrac{1}{2}(E_{\mathrm{cr}}-E_{\mathrm{tr}}). \tag{2D}$$

In the generation of heat in the cable, the squares of the resultant root-mean-square voltage and current are important. By plane geometry, the diagrams show that

$$E^2=E_a^2+E_b^2+2E_aE_b\cos\psi, \tag{3A}$$

$$I^2=I_a^2+I_b^2-2I_aI_b\cos\psi, \tag{3B}$$

where $\psi=4\pi x/\lambda$.

The power flowing toward the load is E_a^2/Z_0, and that flowing away from the load is E_b^2/Z_0. Then, the net power delivered to the load is

$$W=\frac{E_a^2-E_b^2}{Z_0}=\frac{E_{\mathrm{cr}}E_{\mathrm{tr}}}{Z_0}. \tag{4A}$$

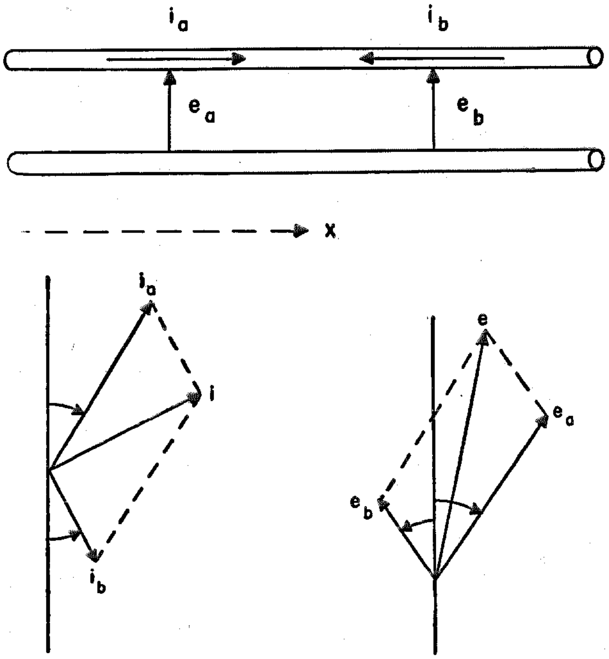

Fig. 1—Voltages and currents on transmission line: $\psi=4\pi x/\lambda$, where x is measured from a voltage maximum. Each angle marked with an arrow is $\psi/2$. Time is $t=0$.

If the line were "flat," then $E_a=E_{\mathrm{flat}}$ and $E_b=0$, so

$$W=\frac{E_{\mathrm{flat}}^2}{Z_0}. \tag{4B}$$

For the same power in (4A) and (4B), there results

$$\left.\begin{array}{l} E_{\mathrm{cr}}=E_{\mathrm{flat}}\sqrt{(\mathrm{swr})} \\ \text{and} \\ E_{\mathrm{tr}}=\dfrac{E_{\mathrm{flat}}}{\sqrt{(\mathrm{swr})}}, \end{array}\right\} \tag{4C}$$

where $E_{\mathrm{cr}}/E_{\mathrm{tr}}=$ standing-wave ratio. Whence

$$\left.\begin{array}{l} E_a=\dfrac{E_{\mathrm{flat}}}{2}\left[\sqrt{(\mathrm{swr})}+\dfrac{1}{\sqrt{(\mathrm{swr})}}\right], \\ \text{and} \\ E_b=\dfrac{E_{\mathrm{flat}}}{2}\left[\sqrt{(\mathrm{swr})}-\dfrac{1}{\sqrt{(\mathrm{swr})}}\right], \end{array}\right\} \tag{4D}$$

and similarly for I_a and I_b.

The power loss in the cable originates in the conductors and dielectric, these losses being proportional to I^2 and E^2, respectively. The average total power loss over a half wavelength is proportional to $E_a^2+E_b^2$. Consequently, for the same power to be delivered to the load when the cable has standing waves as for a "flat" cable,

$$\frac{\text{Power loss}}{\text{Loss for ``flat'' cable}} \approx \frac{1}{2}\left[(\text{swr}) + \frac{1}{(\text{swr})}\right]. \quad \text{(4E)}$$

when the loss is small. It will be shown that the average temperature rise of each conductor, and that of the dielectric, is proportional to the power loss. Thus, the average rise of a cable with standing waves is to that of a "flat" cable as the above ratio of power losses. For example, compared to the attenuation and average temperature rise of a "flat" cable, these values must be multiplied by 1.25 for (swr) = 2.0, and by 2.125 for (swr) = 4.0.

Fig. 2 shows the voltage and current waves E and I when (swr) = 4.0, and also the squares, E^2 and I^2. Note that whereas $E^2_{\text{flat}} = 1.00$, the maximum E^2 is 4.0, and similarly for I^2. Thus, there are localized power losses of four times the uniformly distributed loss in a "flat" cable.

The rate of generation of heat on the surface of the inner conductor, in calories per second per unit length, is

$$q_1 = 0.24 I^2 R_1, \quad \text{(5A)}$$

and for the inside surface of the outer conductor,

$$q_2 = 0.24 I^2 R_2, \quad \text{(5B)}$$

where R_1 and R_2 are the resistance per unit length of the respective conductors at the frequency $\omega/2\pi$.

For computing the heat generated in the dielectric, the cable is considered to be a capacitor with coaxial cylindrical plates of radius r_1 for the outside surface of the inner conductor, and r_2 for the inside surface of the outer conductor. In Fig. 3, for a coaxial cylindrical shell or element in the dielectric, the capacitance per unit length is

$$0.0885\epsilon \frac{2\pi r}{dr} \times 10^{-12} \text{ farads,}$$

where ϵ = dielectric constant. With a radio-frequency current I_ϵ flowing through the dielectric between inner and outer conductors, the

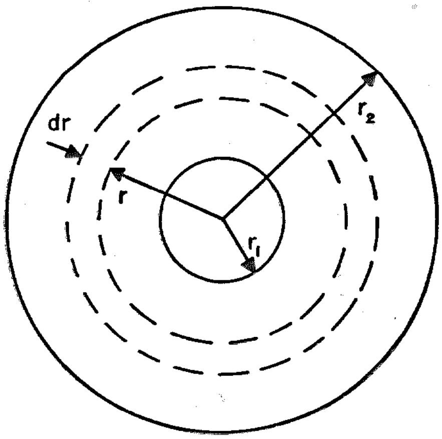

Fig. 3—A section of the dielectric.

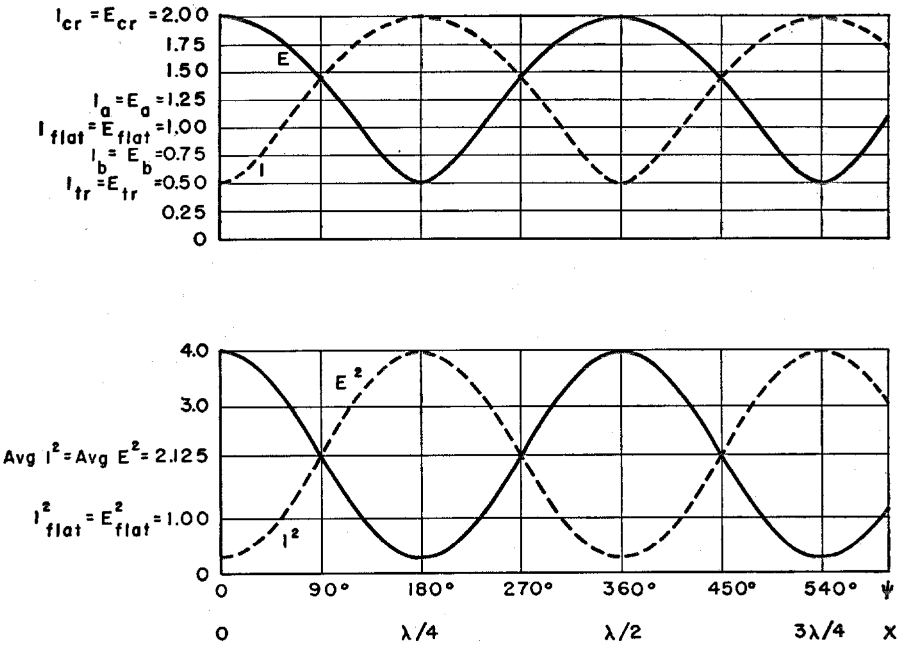

Fig. 2—Standing waves E and I (upper chart) and their squares E^2 and I^2 (lower chart) for a cable with standing-wave ratio = 4.0. Arbitrary units are used which make $E_{\text{flat}} = 1.00$ and $I_{\text{flat}} = 1.00$ for same power delivered to the load.

voltage across the cylindrical shell will be

$$dE=\frac{I_\epsilon}{\omega}\frac{10^{12}\,dr}{0.0885\epsilon\times 2\pi r}.$$

Integrating between the limits r_1 and r_2

$$E=\frac{I_\epsilon}{\omega}\frac{10^{12}\log(r_2/r_1)}{2\pi\times 0.0885\epsilon}=\frac{I_\epsilon}{\omega C},$$

where C is the capacitance in farads per unit length of the coaxial cable. The rate of heat generation in the cylindrical shell in calories per second per unit length is

$$dq_p=0.24I_\epsilon(\mathrm{pf})dE=\frac{0.24(\mathrm{pf})\omega CE^2}{\log(r_2/r_1)}\frac{dr}{r}, \qquad (6)$$

this being an incremental or infinitesimal quantity because the volume of the shell is infinitesimal. If this quantity is integrated between r_1 and r_2, the total heat generated in a unit length of the dielectric is

$$q_p=0.24(\mathrm{pf})\omega CE^2. \qquad (7)$$

3. *Thermal Problem*

It will be observed in the discussion below that the equations for heat flow are linear, since they involve only constants and the first power of the temperature u, and of its derivatives, and no products of such terms. Therefore, the principal of superposition may be applied, just as in electrical problems. That is, if a source of heat is divided into two parts, or if there are several sources of heat, each source may be treated separately and the resulting temperature rises may then be added.

Heat flow is completely analogous to electric current; temperature difference, to potential difference; and thermal conductivity, to electrical conductivity.

In most practical cases, heat may be considered to flow only axially along the inner conductor. Although it flows both axially and radially in the outer conductor, usually the radial temperature gradient is negligibly small. This is due to the small radial dimensions of the conductor compared to the thermal wavelength of the cable (which is half the electrical wavelength). In a 7-strand inner conductor, however, each of the 6 outer stands will be heated equally, and uniformly over their cross section at any point, because of their small diameter and high conductivity. The central conductor, or 7th strand, is heated only by conduction (which may be relatively poor) from the outer conductors. It thus forms a shunt path for the heat, and the over-all conductivity will be somewhat poorer than the ideal.

In the dielectric, on the other hand, the principal flow of heat is radial. The axial flow is usually much less than that in the copper, because the thermal conductivity of the dielectric is usually much lower than that of copper. The flow begins to deviate appreciably from the radial direction only when the thickness of the dielectric becomes an appreciable fraction of a thermal wavelength. However, in this case the radial thickness is also an appreciable fraction of an electrical wavelength, and the cable is in the transition stage between a transmission line and a wave guide. The assumption that the electric flux is radial no longer holds. In the thermal problem, the heat flow is normal to the isothermal (or equal-temperature) surfaces, and in the electrical problem, the electric flux is normal to the equipotential surfaces. While the principal heat flow in the dielectric is radial, and most of the equations are derived on this basis, correction terms are developed later to take care of the axial flow. Generally, the flow may be considered to have circular symmetry about the axis. This will not be quite true if the cable is attached to a surface so the dissipation at the outside jacket does not have circular uniformity, or if there is uneven heating of the surface by the sun or other source of radiation. Gas-filled lines may have the circular symmetry as well as the axial flow appreciably modified by convection.

4. *Equations of Heat Flow*

In the conductor of Fig. 4, which may be solid, tubular, or stranded, but of equivalent thermal conductivity k_c and cross-sectional area

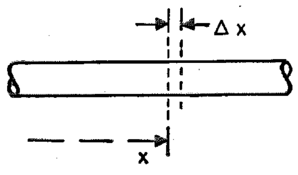

Fig. 4—A portion of the conductor.

A, the heat may usually be considered to be flowing axially, as stated above. At a plane normal to the axis, or x coordinate, heat flow per second in the direction of increasing x is

$$q=-k_c A\frac{du}{dx}, \qquad (8A)$$

where du/dx is the temperature gradient. At $x+\Delta x$, the temperature gradient is $\frac{du}{dx}+\frac{d^2u}{dx^2}\Delta x$. The heat flow entering the element Δx through the conductor is the sum of that flowing into it across the surfaces x and $x+\Delta x$, or

$$\Delta q=k_c A\frac{d^2u}{dx^2}\Delta x. \qquad (8B)$$

This axial heat flow into the element, plus the heat generated therein, is equal to the radial flow out of its surface under steady-state conditions. In the case where equilibrium or steady-state conditions have not been established, there must be added the rate of heat storage within the element. This latter is

$$\Delta q=s_c m_c A\frac{du}{dt}\Delta x, \qquad (9)$$

where s_c is the specific heat or thermal capacity, and m_c is the mass of the conductor per unit volume (density).

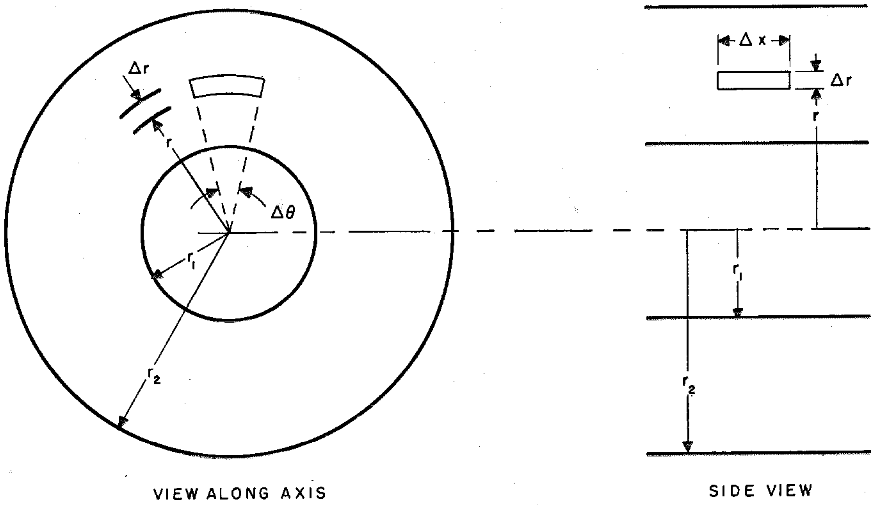

Fig. 5—An element in the dielectric.

In the dielectric of Fig. 5, the heat flow at a point with coordinates r, θ, and x, may be resolved into three mutually perpendicular directions. The net flow along the axial direction into the element shown is

$$k_p r\Delta\theta\Delta r\frac{\partial^2 u_p}{\partial x^2}\Delta x. \qquad (10)$$

The heat flowing into the element radially at the radius r is

$$-k_p r\Delta\theta\Delta x\frac{\partial u_p}{\partial r}, \qquad (11A)$$

and the flow into the element at $r+\Delta r$ is

$$k_p(r+\Delta r)\Delta\theta\Delta x\left(\frac{\partial u_p}{\partial r}+\frac{\partial^2 u_p}{\partial r^2}\Delta r\right).$$

Then the net radial flow into the element, neglecting the small second-order quantities is

$$k_p r\Delta\theta\Delta x\Delta r\left(\frac{\partial^2 u_p}{\partial r^2}+\frac{1}{r}\frac{\partial u_p}{\partial r}\right). \qquad (11B)$$

In the third perpendicular direction, the flow into the element is

$$-k_p\Delta r\Delta x\frac{\partial u_p}{r\partial\theta}+k_p\Delta r\Delta x\left(\frac{\partial u_p}{r\partial\theta}+\frac{\partial}{\partial\theta}\frac{\partial u_p}{r\partial\theta}\Delta\theta\right)$$
$$=k_p\Delta r\Delta x\frac{1}{r}\frac{\partial^2 u_p}{\partial\theta^2}\Delta\theta. \qquad (12)$$

The total heat flow into the element is then

$$\Delta q=k_p r\left(\frac{\partial^2 u_p}{\partial r^2}+\frac{1}{r}\frac{\partial u_p}{\partial r}+\frac{\partial^2 u_p}{\partial x^2}+\frac{1}{r^2}\frac{\partial^2 u_p}{\partial\theta^2}\right)\Delta r\Delta x\Delta\theta. \qquad (13A)$$

The quantity in parenthesis is the Laplacian, or the divergence of the gradient of u, expressed in cylindrical coordinates. In the case of circular symmetry, $\partial^2 u_p/\partial\theta^2=0$, and $\Delta\theta=2\pi$. Also, in the event of negligible axial flow, $\partial^2 u_p/\partial x^2=0$ for practical purposes, and

$$\Delta q=k_p 2\pi r\left(\frac{\partial^2 u_p}{\partial r^2}+\frac{1}{r}\frac{\partial u_p}{\partial r}\right)\Delta r\Delta x. \qquad (13B)$$

This heat flow into the element plus that generated therein by dielectric heating is equal to zero in the steady state, while in the transient condition,

$$\Delta q = s_p m_p \frac{du_p}{dt} r\Delta\theta\Delta r\Delta x. \tag{14}$$

At the boundary between dielectric and inner conductor, the heat flow into the dielectric is, for circular symmetry,

$$-k_p 2\pi r_1 \Delta x \frac{\partial u_p}{\partial r}\Big]_{r_1}, \tag{15A}$$

and at the outer conductor,

$$k_p 2\pi r_2 \Delta x \frac{\partial u_p}{\partial r}\Big]_{r_2}. \tag{15B}$$

When no heat is generated in the dielectric, and under steady-state conditions with heat assumed to be flowing only radially through the dielectric from inner to outer conductor, the flow per unit axial length has the value, independent of r,

$$q = -2\pi k_p r \frac{du_p}{dr}.$$

Separating the variables u and r, and integrating between r_1 and any radius r, such as r_2,

$$\left.\begin{aligned} q &= \frac{2\pi k_p}{\log (r/r_1)}(u_1 - u_p), \\ &= \frac{2\pi k_p}{\log (r_2/r_1)}(u_1 - u_2). \end{aligned}\right\} \tag{16}$$

It follows that the temperature in the dielectric is a function of r:

$$u_p = u_1 + (u_2 - u_1)\frac{\log (r/r_1)}{\log (r_2/r_1)}, \tag{17}$$

when no heat is generated therein, and the flow is radial. This is the same law as that referring to the electrical potential as a function of r.

This result can also be derived from (13B). Since the heat flow is in equilibrium, $\Delta q = 0$, and

$$\frac{d^2u_p}{dr^2} + \frac{1}{r}\frac{du_p}{dr} = 0.$$

A solution of this differential equation is

$$u_p = a + b \log \frac{r}{r_1}.$$

Equation (17) results when the following boundary conditions are submitted:

$$u_p = u_1 \quad \text{at} \quad r = r_1, \quad \text{and} \quad u_p = u_2 \quad \text{at} \quad r = r_2.$$

Phenomena taking place at the surface of the cable include the dissipation of the heat generated internally and the heating of the cable by the external surroundings; radiation from the sun, etc. If the dissipation of heat is the only matter of concern, then, per unit length,

$$\Delta q = \mathcal{E}_s 2\pi r_s \Delta x (u_2 - u_0),$$

where $\mathcal{E}_s$ is the surface emissivity, a constant depending on the nature of the surface of the cable and on the medium in which it is immersed. The radius of the surface is r_s, while u_2 is the actual temperature of the outer conductor, and u_0 the ambient temperature of the medium. If the cable has a jacket of conductivity k_s, inner radius r_3, outer radius r_s, and $\mathcal{E}_s$ is the surface emissivity of the jacket, then the dissipation per unit length is

$$q = \frac{2\pi k_s}{\log (r_s/r_3)}(u_2 - u_3) = 2\pi r_s \mathcal{E}_s (u_3 - u_0),$$

where u_3 is the actual temperature of the outside of the jacket. This may be written

$$q = 2\pi r_s \mathcal{E} (u_2 - u_0), \tag{18A}$$

where $\mathcal{E}$ is now the effective total emissivity of the outer conductor to the surrounding medium, this being found by the relationship

$$\frac{1}{\mathcal{E}} = \frac{r_s}{k_s} \log \frac{r_s}{r_3} + \frac{1}{\mathcal{E}_s}, \tag{18B}$$

which is analogous to two resistors in series.

5. *Heat Generated on Inner Conductor*

According to (3B) and (5A), the heat generated on a section of the inner conductor of unit length is

$$\left.\begin{aligned} q_1 &= 0.24 R_1 (I_a^2 + I_b^2 - 2 I_a I_b \cos \psi) \\ &= q_1' - q_1'' \cos \psi, \end{aligned}\right\} \tag{19}$$

which indicates a component uniformly distributed along the conductor, and a sinusoidally varying component. These two components may be treated separately, since the equations of heat flow are linear.

Taking the uniform part of the current-squared distribution, $I_a^2+I_b^2$, it is evident that for a cable in a uniform medium of constant ambient temperature, or for a part of the cable at some distance from a variation in the medium, the temperature of each conductor will be independent of x, and there will be no heat flow parallel to the axis in either the dielectric or the conductors. The dielectric becomes a simple medium with heat conduction in accordance with (16). At the surface, by (18A),

$$q=2\pi r_s\mathcal{E}(u_2-u_0) \tag{20}$$

per unit length. The rates q and q_1' in (19) and (20) are equal to each other, being the rate of heat generation per unit length of the inner conductor;

$$q_1'=0.24(I_a^2+I_b^2)R_1.$$

Therefore, the temperature rise of the outer conductor above ambient is

$$u_{21}'=\frac{q_1'}{2\pi r_s\mathcal{E}}. \tag{21A}$$

By (16), the temperature rise of the inner conductor is

$$u_{11}'=u_{21}'+\frac{q_1'}{2\pi k_p}\log\frac{r_2}{r_1}. \tag{21B}$$

The variation of temperature with radius in the dielectric is given by (17).

The part of (19) which is a function of x is

$$-q_1''\cos\psi=-0.48R_1I_aI_b\cos\psi, \tag{22A}$$

which is the rate of generation of heat per unit length. This is equivalent to a sinusoidal heating and cooling of the conductor with a wavelength $\lambda/2$, since $\psi=4\pi x/\lambda$. Maximum heating occurs at the current-crest point of the standing waves on the cable, and maximum apparent cooling (relative to the heating caused by q_1') at the current trough.

In a uniform medium, this can be expected to give a periodic temperature variation on the inner conductor,

$$u_{11}''=\textstyle\sum_n U_{11}\cos(n\psi+\phi_n), \tag{22B}$$

where n is any positive integer. The temperatures of the dielectric and of the outer conductor will vary periodically with the same period, and the temperatures therein can be expressed as summations similar to that for u_{11}''.

If the conductor were not in a cable, but isolated and simply immersed in a medium of ambient temperature u_0, then by (8B) and (18A),

$$k_cA_1\frac{d^2u_1}{dx^2}+2\pi r_s\mathcal{E}(u_0-u_1)-q_1''\cos\psi=0. \tag{23A}$$

That is, the heat generated in a section Δx of the conductor, plus that flowing along the conductor into the section, plus that from the surrounding medium is equal to zero.

It may be pointed out that all heating and apparent cooling of the conductor causes a temperature change with respect to the ambient temperature and independent of the ambient. This follows from the linear nature of the equations, as pointed out before. In the present case, as a result of the ambient acting alone, there is a temperature distribution u_{10} in the conductor;

$$k_cA_1\frac{d^2u_{10}}{dx^2}-2\pi r_s\mathcal{E}(u_{10}-u_0)=0,$$

where u_0 may be a function of x. The temperature u_{11}, because of the electrical heating of the conductor in the ambient temperature $u_0=0$, is

$$k_cA_1\frac{d^2u_{11}}{dx^2}-2\pi r_s\mathcal{E}u_{11}-q_1''\cos\psi=0.$$

Adding,

$$k_cA_1\frac{d^2(u_{10}+u_{11})}{dx^2}-2\pi r_s\mathcal{E}(u_{10}+u_{11}-u_0)$$
$$-q_1''\cos\psi=0. \tag{23B}$$

By substituting $u_{10}+u_{11}=u_1$, equation (23A) results, and so the two effects may be treated separately and the resulting temperature rise due to the heating added to the temperature distribution caused by the ambient. Throughout the investigation below it will be customary to set $u_0=0$ in the equations and then to add that rise due to the ambient (the latter is U_0 for some distance in the vicinity of the point in question when the ambient is constant) to the final rise.

Solving (23A) with $u_0=0$, the complementary function is the solution of

$$k_cA_1\frac{d^2u_1}{dx^2}-2\pi r_s\mathcal{E}u_1=0,$$

which is

$$u_1=G\epsilon^{\alpha x}+H\epsilon^{-\alpha x}, \tag{24A}$$

where

$$\alpha=\left(\frac{2\pi r_s}{A_1}\frac{k_s}{k_c}\right)^{1/2}, \tag{24B}$$

and G and H are arbitrary constants. This would give the "space transient" condition in the vicinity of a change of medium, or termination of the cable in a fitting or open end. As a uniform medium is stipulated above, $G=0$ and $H=0$ in the present case.

For the particular integral of (23A), change the variable:

$$\frac{d^2u_1}{dx^2}=\left(\frac{d\psi}{dx}\right)^2\frac{d^2u_1}{d\psi^2}=\left(\frac{4\pi}{\lambda}\right)^2\frac{d^2u_1}{d\psi^2}. \tag{25A}$$

Then the temperature distribution u_1'' along the conductor is determined by

$$k_cA_1\left(\frac{4\pi}{\lambda}\right)^2\frac{d^2u_1''}{d\psi^2}-2\pi r_s\mathcal{E}u_1''=q_1''\cos\psi. \tag{25B}$$

The particular integral is of the form (22B) which may be verified by substitution. Since any value may be assigned to ψ, the equation must be satisfied independently for each argument $n\psi$. Then ${}_0U_1=0$ and all values of ${}_nU_1=0$ except for $n=1$, which may be written U_1. Therefore,

$$u_1''=U_1\cos\psi=\frac{-q_1''\cos\psi}{k_cA_1\left(\dfrac{4\pi}{\lambda}\right)^2+2\pi r_s\mathcal{E}}. \tag{26}$$

To this must be added the ambient u_0 and the temperature rise u_1' resulting from the constant component q_1' of the generated heat.

In the complete cable, two simultaneous equations must be satisfied, one for each conductor. These follow from (16) and (23A), neglecting at present any axial flow in the dielectric

$$\begin{cases} k_cA_1\dfrac{d^2u_{11}''}{dx^2}+\dfrac{2\pi k_p}{\log(r_2/r_1)}(u_{21}''-u_{11}'') \\ \qquad\qquad -q_1''\cos\psi=0, & (27A) \\ k_cA_2\dfrac{d^2u_{21}''}{dx^2}+\dfrac{2\pi k_p}{\log(r_2/r_1)}(u_{11}''-u_{21}'') \\ \qquad\qquad -2\pi r_s\mathcal{E}u_{21}''=0. & (27B) \end{cases}$$

Proceeding as before, the integration constants of the complementary function or "space transient" equal zero, and there remains the particular integral

$$u_{21}''=U_{21}\cos\psi=\frac{-q_1''\cos\psi}{2\pi r_s\mathcal{E}+\left(\dfrac{4\pi}{\lambda}\right)^2k_c(A_1+A_2)+\dfrac{k_cA_1}{2\pi k_p}\left(\dfrac{4\pi}{\lambda}\right)^2\left[k_cA_2\left(\dfrac{4\pi}{\lambda}\right)^2+2\pi r_s\mathcal{E}\right]\log\dfrac{r_2}{r_1}}, \tag{28A}$$

$$u_{11}''=U_{11}\cos\psi, \tag{28B}$$

where

$$U_{11}=U_{21}\left[1+\frac{k_cA_2\left(\dfrac{4\pi}{\lambda}\right)^2+2\pi r_s\mathcal{E}}{2\pi k_p}\log\frac{r_2}{r_1}\right].$$

The temperature at any point in the dielectric is found by means of (17).

6. *Heat Generated on Outer Conductor*

For heat generated on the outer conductor, the steady-state temperature rise above ambient resulting from the constant component $q_2'=0.24(I_a^2+I_b^2)R_2$ of (3B) and (5B) is the same for both the inner and outer conductors:

$$u_{12}'=u_{22}'=\frac{q_2'}{2\pi r_s\mathcal{E}}. \tag{29}$$

For the component of the heat that is a function of x or ψ, $q_2''=0.48I_aI_b\cos\psi$, the equations are, where axial flow in the dielectric is negligible,

$$\begin{cases} k_cA_1\dfrac{d^2u_{12}''}{dx^2}+\dfrac{2\pi k_p}{\log(r_2/r_1)}(u_{22}''-u_{12}'')=0, & (30A) \\ k_cA_2\dfrac{d^2u_{22}''}{dx^2}+\dfrac{2\pi k_p}{\log(r_2/r_1)}(u_{12}''-u_{22}'') \\ \qquad\qquad -2\pi r_s\mathcal{E}u_{22}''-q_2''\cos\psi=0. & (30B) \end{cases}$$

Solving:

$$u_{22}''=U_{22}\cos\psi, \tag{31A}$$

$$u_{12}''=U_{12}\cos\psi, \tag{31B}$$

where

$$U_{22}=\frac{-q_2''}{2\pi r_s\mathcal{E}+k_cA_2\left(\dfrac{4\pi}{\lambda}\right)^2+\dfrac{1}{\dfrac{\log(r_2/r_1)}{2\pi k_p}+\dfrac{1}{k_cA_1}\left(\dfrac{\lambda}{4\pi}\right)^2}}, \tag{31C}$$

and

$$U_{12}=U_{22}\frac{2\pi k_p}{\log(r_2/r_1)}\,\frac{1}{k_cA_1\left(\frac{4\pi}{\lambda}\right)^2+\frac{2\pi k_p}{\log(r_2/r_1)}}=\frac{U_{22}}{1+\frac{k_cA_1}{2\pi k_p}\left(\frac{4\pi}{\lambda}\right)^2\log\frac{r_2}{r_1}}. \tag{31D}$$

7. *Heat Generated in Dielectric*

When the heat conductivity of the dielectric is small compared to that of the conductor material, and for usual cross sections of cable, the axial heat flow in the dielectric can be neglected to a first approximation, as was done in the cases treated above. Then by (3A), (6), and (13B), the differential equation for the temperature distribution in the dielectric is

$$r^2\frac{d^2u_p}{dr^2}+r\frac{du_p}{dr}+\frac{q_p}{2\pi k_p\log(r_2/r_1)}=0, \tag{32A}$$

where

$$\begin{aligned}q_p&=0.24(\mathrm{pf})\omega C(E_a^2+E_b^2+2E_aE_b\cos\psi) \qquad (32\mathrm{B})\\&=q_p'+q_p''\cos\psi.\end{aligned}$$

A solution of (32A) is

$$u_p=a+b\log\frac{r}{r_1}-\frac{q_p}{4\pi k_p\log\frac{r_2}{r_1}}\left(\log\frac{r}{r_1}\right)^2. \tag{33A}$$

The boundary conditions are $u_p=u_1$ at $r=r_1$ and $u_p=u_2$ at $r=r_2$, whence

$$\begin{aligned}u_p=u_1+&\left(\frac{u_2-u_1}{\log(r_2/r_1)}+\frac{q_p}{4\pi k_p}\right)\log\frac{r}{r_1}\\&-\frac{q_p}{4\pi k_p\log(r_2/r_1)}\left(\log\frac{r}{r_1}\right)^2. \qquad (33\mathrm{B})\end{aligned}$$

It is to be noted that (33B) is general; it includes the temperature distribution resulting from heat generated externally to the dielectric, as well as that generated internally. It reduces to (17) when $q_p=0$, and to (37C) when $q_p''=0$, provided that only the heat generated in the dielectric is considered.

This result may be drived also by simple integration. The heat generated in a section of the dielectric of unit axial length, between the limits r_1 and r is determined by integrating (6), and is found to be:

$$\frac{0.24(\mathrm{pf})\omega CE^2}{\log(r_2/r_1)}\log\frac{r}{r_1}=\frac{q_p}{\log(r_2/r_1)}\log\frac{r}{r_1}.$$

To this must be added h_1 the heat flowing into the section from the conductor at radius r_1, giving the total heat flowing outward from the section at radius r. By (11A), the sum of these is equal to

$$-2\pi rk_p\frac{du_p}{dr}.$$

The result of this equality is

$$-2\pi rk_p\frac{du_p}{dr}=h_1+\frac{q_p}{\log(r_2/r_1)}\log\frac{r}{r_1}. \tag{34}$$

Separating variables, and noting that

$$\int du_p=u_p,\quad \int\frac{dr}{r}=\log r,$$

and

$$\int\frac{1}{r}\log r\,dr=\frac{1}{2}(\log r)^2$$

plus an arbitrary constant, there results a solution which can be written in the form (33A). Dividing (34) through by r and differentiating yields the differential equation (32A).

For the distribution of temperature along the conductors resulting from heat generated in the dielectric, it is noted that in the steady state the net heat flow into an element Δx of the conductor is zero. For the inner conductor, by (8B) and (15A), and for the outer conductor by (8B), (15B), and (18A), letting $u_0=0$:

$$k_cA_1\frac{\partial^2u_{1p}}{\partial x^2}+2\pi r_1k_p\frac{\partial u_p}{\partial r}\bigg]_{r_1}=0, \tag{35A}$$

$$k_cA_2\frac{\partial^2u_{2p}}{\partial x^2}-2\pi r_2k_p\frac{\partial u_p}{\partial r}\bigg]_{r_2}-2\pi r_s\mathcal{E}u_{2p}=0. \tag{35B}$$

Differentiating (33A) or (33B),

$$\frac{\partial u_p}{\partial r}=\frac{b}{r}-\frac{q_p}{2\pi rk_p\log(r_2/r_1)}\log\frac{r}{r_1}. \tag{35C}$$

Which, substituted in (35A) and (35B), gives

$$k_cA_1\frac{\partial^2u_{1p}}{\partial x^2}+2\pi k_p\frac{u_{2p}-u_{1p}}{\log(r_2/r_1)}+\frac{q_p}{2}=0, \tag{36A}$$

$$\begin{aligned}k_cA_2\frac{\partial^2u_{2p}}{\partial x^2}+2\pi k_p\frac{u_{1p}-u_{2p}}{\log(r_2/r_1)}+\frac{q_p}{2}&\\-2\pi r_s\mathcal{E}u_{2p}=0.&\qquad(36\mathrm{B})\end{aligned}$$

Note that the heat entering the dielectric, the negative of the second plus third terms of (36A), plus the heat q_p generated therein is equal to that leaving the dielectric, which is the second plus third terms of (36B).

Taking first the constant part q_p' of (32B), a reasonable supposition is that u_{1p} and u_{2p} are independent of x, as was the case for the constant part of the heat generated on the conductors. It follows that $\partial u_{1p}/\partial x=0$ and $\partial u_{2p}/\partial x=0$. From (35A), or from the fact that conditions are in equilibrium (no heat flow into or out of the inner conductor), it follows that

$$\left.\frac{\partial u_p}{\partial r}\right]_{r_1}=0.$$

Whence, from (36A)

$$u_{1p}'=u_{2p}'+\frac{q_p'}{4\pi k_p}\log\frac{r_2}{r_1}, \qquad (37A)$$

and from (36A) and (36B), or from elementary considerations

$$u_{2p}'=\frac{q_p'}{2\pi r_s\mathcal{E}}. \qquad (37B)$$

When (37A) is substituted in (33B), it is found that the temperature distribution due to q_p' acting alone is

$$u_{pp}'=u_{1p}'-\frac{q_p'}{4\pi k_p\log(r_2/r_1)}\left(\log\frac{r}{r_1}\right)^2. \qquad (37C)$$

When $q_p=0$, (36A) and (36B) naturally reduce to the same equations as do (27A) and (27B), and also (30A) and (30B) when $q_1=0$ or $q_2=0$:

$$\frac{d^2u_1}{dx^2}+\frac{2\pi k_p}{k_cA_1\log(r_2/r_1)}(u_2-u_1)=0, \qquad (38A)$$

$$\frac{d^2u_2}{dx^2}+\frac{2\pi k_p}{k_cA_2\log(r_2/r_1)}(u_1-u_2)-\frac{2\pi r_s\mathcal{E}}{k_cA_2}u_2=0. \qquad (38B)$$

This gives the "space transient" condition, similar to (24A), from which may be determined the temperatures in the cable resulting from a change of medium or end effects, such as the presence of fittings.

The variable part of q_p, which is $q_p''\cos\psi$, when substituted in (36A) and (36B), gives equations of a type similar to (27A) and (27B) and also (30A) and (30B). The solution is

$$u_{2p}''=U_{2p}\cos\psi, \qquad (39A)$$

$$u_{1p}''=U_{1p}\cos\psi. \qquad (39B)$$

It can be readily shown that for $q_p=q_p''\cos\psi$, (36A) and (36B) are not satisfied by any terms $u_{2p}''={}_nU_{2p}\cos n\psi$, except (39A), in which $n=1$. Similarly for u_{1p}''.

The coefficients U_{1p} and U_{2p} are determined by substituting (39) into (36A) and (36B):

$$-\left[k_cA_1\left(\frac{4\pi}{\lambda}\right)^2+\frac{2\pi k_p}{\log(r_2/r_1)}\right]U_{1p}+\frac{2\pi k_p}{\log(r_2/r_1)}U_{2p}=-\frac{q_p''}{2}, \qquad (40A)$$

$$\frac{2\pi k_p}{\log(r_2/r_1)}U_{1p}-\left[k_cA_2\left(\frac{4\pi}{\lambda}\right)^2+\frac{2\pi k_p}{\log(r_2/r_1)}+2\pi r_s\mathcal{E}\right]U_{2p}=-\frac{q_p''}{2}. \qquad (40B)$$

The simplest means of solving (40A) and (40B) is to substitute the numerical values in the coefficients from the dimensions and constants of the cable. Then solve the resulting linear simultaneous equations for U_{1p} and U_{2p}.

The temperature distribution in the dielectric is given by (33B).

8. *Example*

As an example of the practical application of the equations developed in this paper, consider a Type RG-19/U cable delivering 5 kilowatts continuously to a load, the frequency being 100 megacycles per second. Two cases will be considered: A. where the cable is "flat," or standing-wave ratio = 1; B. where the standing-wave ratio is 2.0.

This cable is made up as follows:

Inner conductor: solid plain copper, 0.250-inch diameter.

Cable core: solid polyethylene, 0.910-inch outside diameter.

Outer conductor: single-layer braid, 48×9 30-gauge (Brown and Sharpe) plain copper wires. Maximum (allowable but not necessarily actual) diameter is 0.990 inch.

Jacket: synthetic resin, outside diameter 1.120 inches.

The specified nominal characteristics are:

Surge impedance = 52 ohms.
Capacitance = 29.5 micromicrofarads per foot.

Attenuation = 0.68 decibel per 100 feet at 100 megacycles.

Then, the various constants of the cable are:

$r_1 = 0.318$ centimeter.
$r_2 = 1.16$ centimeters.
$r_3 = 1.21$ centimeters (estimated average value).
$r_s = 1.42$ centimeters.
$A_1 = 0.317$ square centimeter.
$A_2 = 0.220$ square centimeter.
$k_c = 1.0$ calorie per second per square centimeter for one degree centigrade per centimeter.
$k_p = 8 \times 10^{-4}$ calorie per second per square centimeter for one degree centigrade per centimeter.
$k_s = 4 \times 10^{-4}$ calorie per second per square centimeter for one degree centigrade per centimeter (estimated).
$\mathcal{E}_s = 3 \times 10^{-4}$ calorie per second per square centimeter for one degree centigrade temperature difference (estimated; depends upon conditions of installation).
$\epsilon = 2.3$ (dielectric constant of core).
$(\mathrm{pf}) = 4 \times 10^{-4}$.
$C = 0.97 \times 10^{-12}$ farad per centimeter.
$Z_0 = 52$ ohms.

From these constants other derived parameters are found:

$\mathcal{E} = 2.6 \times 10^{-4}$ calorie per second per square centimeter for one degree centigrade temperature difference; by equation (18B).
$R_1 = 1.48 \times 10^{-3}$ ohm per centimeter.
$R_2 = 0.525 \times 10^{-3}$ ohm per centimeter.

The values of R_1 and R_2 are taken to be slightly higher than the ideal theoretical values for 100 megacycles, but together with the power-factor value listed, when used to calculate[2] the attenuation of the cable, result in the specified value of 0.68 decibel per 100 feet.

For the given frequency, power, and standing-wave ratio, the following quantities are found:

$\lambda = 198$ centimeters.

A. For the "flat" line, standing-wave ratio = 1:

$E_{\text{flat}} = 510$ root-mean-square volts. (4B)
$I_{\text{flat}} = 9.80$ root-mean-square amperes. (1E)
$q_1 = 34.1 \times 10^{-3}$ calorie per second per centimeter. (5A)
$q_2 = 12.1 \times 10^{-3}$ calorie per second per centimeter. (5B)
$q_p = 15.2 \times 10^{-3}$ calorie per second per centimeter. (7)

B. For the line with standing-wave ratio = 2.0:

$E_a = 540$ root-mean-square volts. (4D)
$E_b = 180$ root-mean-square volts. (4D)
$I_a = 10.4$ root-mean-square amperes. (1E)
$I_b = 3.46$ root-mean-square amperes. (1E)
$q_1' = 42.5 \times 10^{-3}$ calorie per second per centimeter. (19)
$q_1'' = 25.6 \times 10^{-3}$ calorie per second per centimeter. (19)
$q_2' = 15.1 \times 10^{-3}$ calorie per second per centimeter.
$q_2'' = 9.1 \times 10^{-3}$ calorie per second per centimeter.
$q_p' = 19.0 \times 10^{-3}$ calorie per second per centimeter. (32B)
$q_p'' = 11.4 \times 10^{-3}$ calorie per second per centimeter. (32B)

Note that

$$\frac{q_1'}{q_1} = \frac{q_2'}{q_2} = \frac{q_p'}{q_p} = 1.25. \tag{4E}$$

For the "flat" cable carrying 5 kilowatts at 100 megacycles, the total heat generated is $q_1 + q_2 + q_p = 61.4 \times 10^{-3}$ calorie per second per centimeter, as tabulated above. The specified attenuation of the cable at 100 megacycles is 0.68 decibel per 100 feet, which is equal to 0.223×10^{-3} decibel per centimeter. This attenuation acting on 5 kilowatts results in a loss of 0.258 watt per centimeter, or 61.7×10^{-3} calorie per second per centimeter, which closely checks the total heat generated.

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[2] "Reference Data for Radio Engineers," Federal Telephone and Radio Corporation, New York, New York, 2nd Edition; 1946: p. 206.

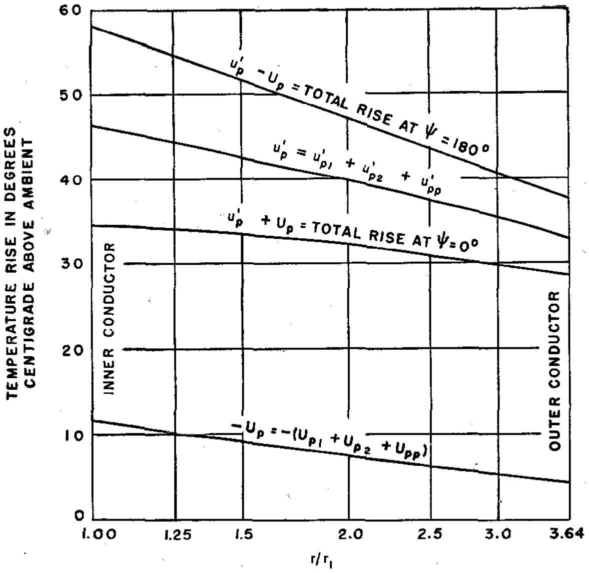

Fig. 6—Temperature rise of the dielectric, computed for a Type RG-19/U cable, carrying 5 kilowatts continuous-wave at 100 megacycles, with standing-wave ratio = 2.0.

For use in applying the formulas, certain quantities which occur several times are here evaluated:

$$\log\frac{r_2}{r_1}=1.29,$$

$$2\pi r_s\mathcal{S}=2.32\times10^{-3},$$

$$\frac{2\pi k_p}{\log\frac{r_2}{r_1}}=3.90\times10^{-3},$$

$$4\pi k_p=10.1\times10^{-3},$$

$$4\pi k_p\log\frac{r_2}{r_1}=13.0\times10^{-3},$$

$$\left(\frac{4\pi}{\lambda}\right)^2=4.05\times10^{-3},$$

$$k_c(A_1+A_2)\left(\frac{4\pi}{\lambda}\right)^2=2.18\times10^{-3},$$

$$k_cA_2\left(\frac{4\pi}{\lambda}\right)^2=0.89\times10^{-3},$$

$$k_cA_1\left(\frac{4\pi}{\lambda}\right)^2=1.28\times10^{-3},$$

$$\frac{k_cA_1}{2\pi k_p}\left(\frac{4\pi}{\lambda}\right)^2=0.258.$$

The principal temperature rises are given in Table I and Fig. 6. The data and curves are the result solely of computation from the specified dimensions and construction of the cable, and the published physical characteristics of its component parts. No experimental tests have been made in this connection to confirm the data. Some of the tabulated temperature rises for the 5-kilowatt power value probably exceed the safe limit for this cable.

The column headed "If k_c were 0" gives hypothetical values which, when compared to the "Actual" column, show the effect of axial flow in the conductors in reducing temperature variation along the cable.

9. Correction for Axial Heat Flow in Dielectric

As explained in the introduction, the heat flow encountered in the dielectric in most practical cable problems is principally radial. All the results so far obtained are based on this assumption, the term $\partial^2u/\partial x^2$ in (13A) always being neglected. With the additional condition of circular symmetry (13B) is obtained. The temperature distribution in the dielectric at any point x is a function of r given by (17) for heat generated in the conductors;[3]

$$u=u_1+(u_2-u_1)\frac{\log\frac{r}{r_1}}{\log\frac{r_2}{r_1}},\qquad(17)$$

and by (33B) for heat generated in the dielectric. For sinusoidal distribution of heating produced by standing waves on the cable, it was found that with axial flow in the dielectric neglected

$$u_2=U_2\cos\psi,\qquad(28A)$$

$$u_1=U_1\cos\psi,\qquad(28B)$$

and therefore by substitution in (17) it follows that

$$u=U\cos\psi,\qquad(41)$$

where U is a function of r only.

The radial temperature distribution (17) is a first approximation to the actual distribution. However, a temperature distribution (41) will produce a resulting flow in the x or ψ direction. The net axial heat flow out of any element of the dielectric (Fig. 5) must be offset by an additional net radial heat flow into the element. This can be accomplished by an increment of radial temperature distribution u_Δ which supplies only the heat required by each element of the dielectric to offset the axial flow resulting from the distribution u. The term u_Δ may be most readily determined on the assumption that it in turn produces negligible axial flow. Then after having determined u_Δ, and as it actually does produce a small axial flow, a second correction term $u_{\Delta\Delta}$ may be determined in the same manner, and so on. The resulting actual temperature at any point becomes the sum of series $u+u_\Delta+u_{\Delta\Delta}$ . . . . , in which the terms may be expected to diminish rapidly in magnitude, and are alternately positive and negative. Thus the series converges rapidly. Each term is a function of r and ψ, and all terms except u are equal to zero for $r=r_1$ and for $r=r_2$.

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[3] In this section, u and U (without subscripts, or with subscripts Δ or $\Delta\Delta$) refer to the temperature *rise* of the dielectric due to heat sources external thereto, the use of subscripts being reduced to the greatest practicable extent.

TABLE I

Type RG-19/U Cable Carrying 5 Kilowatts at 100 Megacycles

Heat Source	Effect	Flat Line (swr) = 1	Standing-Wave Ratio = 2			From Equation
			Part Independent of x	Part Function of x		
				Actual	If k_c Were 0	
Inner Conductor	q_1'	34.1×10^{-3}	42.5×10^{-3}			(19)
	u_{21}'	14.7°	18.3°			(21A)
	u_{11}'	23.4°	29.2°			(21B)
	u_{p1}'		*			(17)
	q_1''			25.6×10^{-3}		(19)
	U_{21}			− 4.6°	−11.0°	(28A)
	U_{11}			−13.0°	−28.5°	(28B)
	U_{p1}			*		
Outer Conductor	q_2'	12.1×10^{-3}	15.1×10^{-3}			
	u_{22}'	5.2°	6.5°			(29)
	u_{12}'	5.2°	6.5°			(29)
	u_{p2}'	5.2°	6.5°			(17)
	q_2''			9.1×10^{-3}		
	U_{22}			−2.2°	−3.9°	(31C)
	U_{12}			−1.6°	−3.9°	(31D)
	U_{p2}			*		(17)
Dielectric	q_p'	15.2×10^{-3}	19.0×10^{-3}			(32B)
	u_{2p}'	6.6°	8.2°			(37B)
	u_{1p}'	8.5°	10.6°			(37A)
	u_{pp}'		*			(37C)
	q_p''			11.4×10^{-3}		(32B)
	U_{2p}			+2.4°	+4.9°	(40A)
	U_{1p}			+2.9°	+6.4°	(40B)
	U_{pp}			*		(33B)
Inner and Outer Conductors and Dielectric, Combined	$q_1'+q_2'+q_p'$	61.4×10^{-3}	76.6×10^{-3}			By Addition
	u_3'	22.8°	28.6°			≈(18A)
	u_2'	26.4°	33.0°			(18A)
	u_1'	37.1°	46.3°			By Addition
	u_p'		(Fig. 6.)			“
	$q_1''+q_2''+q_p''$			46.1×10^{-3}		“
	U_2			− 4.4°		“
	U_1			−11.7°		“
	U_p			(Fig. 6.)		“

* Calculated for Fig. 6, but individual components are not recorded here.

Having determined the series, the simultaneous equations for heat flow in the cable, similar to (27A) and (27B), are set up using (8B), (15A), (15B), and (18A) as was done before. The result will naturally be that the heating of the cable will be slightly less than it appeared to be when axial flow in the dielectric was neglected.

By (13A) and (25A), the equation for heat flow with circular symmetry is

$$\frac{\partial^2 u}{\partial r^2}+\frac{1}{r}\frac{\partial u}{\partial r}=-\left(\frac{4\pi}{\lambda}\right)^2\frac{\partial^2 u}{\partial \psi^2}. \tag{42}$$

If, to a first approximation, the temperature is given by (41) where U is a function of r only, then, by differentiation,

$$-\frac{\partial^2 u}{\partial \psi^2}=U\cos\psi=u, \tag{43}$$

which is proportional to the heat flowing out of an element at coordinates r and ψ. With distribution according to (41) there is a maximum of heat flow out of the element at maximum u, where $\psi=0$. There is no flow through the element at exactly $\psi=0$, the heat flowing equally in both directions. On the other hand, at $\psi=90$ degrees, where $u=0$ for all values of r, there is no net heat flow into the element, but a maximum axial flow through it. Since there is no net flow into an element at $\psi=90$ degrees, the correction term u_Δ is zero at this coordinate. All these results become obvious upon consideration of the physical picture.

The left-hand side of (42) gives the net heat flow radially into the section to compensate for the right side, which is the heat flowing out axially. When (17) is substituted into the left side of (42), it gives zero, because (17) is the solution with the left side set equal to zero. There remains the first correction term u_Δ to be used in the left side. By the reasoning above, since u_Δ is assumed to be small compared to u, it is omitted from the right side of (42). Then by (42), (43), and (17),

$$\begin{aligned}\frac{\partial^2 u_\Delta}{\partial r^2}+\frac{1}{r}\frac{\partial u_\Delta}{\partial r}&=\left(\frac{4\pi}{\lambda}\right)^2 u\\ &=\left(\frac{4\pi}{\lambda}\right)^2\left[u_1+\frac{u_2-u_1}{\log\,(r_2/r_1)}\log\frac{r}{r_1}\right].\end{aligned} \tag{44}$$

As was indicated in connection with (39A) and (39B), since $u=U\cos\psi$, etc., from (28A), (28B), and (41), equation (44) will be satisfied by $u_\Delta=U_\Delta\cos\psi$. Then

$$\frac{d^2U_\Delta}{dr^2}+\frac{1}{r}\frac{dU_\Delta}{dr}=g+h\log\frac{r}{r_1},$$

where g and h are functions of U_1 and U_2, etc., from the right side of (44). Inspection indicates that dU_Δ/dr probably should contain a term $r\log\,(r/r_1)$. Following this through, the solution becomes

$$U_\Delta=a+\frac{r^2}{4}(g-h)+\left(b+h\frac{r^2}{4}\right)\log\frac{r}{r_1}, \tag{45A}$$

where a and b are arbitrary constants determined by the boundary conditions $U_\Delta=0$ at $r=r_1$, and at $r=r_2$. When the constants and boundary conditions are substituted,

$$\begin{aligned}U_\Delta=\left(2\pi\frac{r_1}{\lambda}\right)^2\Bigg\{&\left[U_1+\frac{U_1-U_2}{\log\,(r_2/r_1)}\right]\left[\left(\frac{r}{r_1}\right)^2-1\right]\\ &+\left[\frac{U_1-\left(\frac{r_2}{r_1}\right)^2U_2}{\log\frac{r_2}{r_1}}+\frac{\left(\frac{r_2^2}{r_1^2}-1\right)(U_2-U_1)}{\left(\log\frac{r_2}{r_1}\right)^2}+\left(\frac{r}{r_1}\right)^2\frac{U_2-U_1}{\log\frac{r_2}{r_1}}\right]\log\frac{r}{r_1}\Bigg\}.\end{aligned} \tag{45B}$$

It is to be noted that U_1 and U_2 in (45B) are temperature rises, as indicated by their definition in (28A) and (28B), and not actual temperatures referred to an arbitrary zero.

For most practical cases, the factor $(2\pi r_1/\lambda)^2$ is a small quantity of the order of 0.1 at 3000 megacycles, 0.001 to 0.01 at 600 megacycles, and 0.001 or less at 100 megacycles. That part of U_Δ within the braces is of the same order as the principal temperature term U. Determination of the second correction term $u_{\Delta\Delta}$ by substitution of u_Δ in the right-hand side of (42) instead of u will give a quantity with an over-all multiplying factor of $(2\pi r_1/\lambda)^4$ which will be entirely negligible.

As an example, suppose $(2\pi r_1/\lambda)^2=0.01$, $U_1=45$ degrees, $U_2=15$ degrees, and $r_2/r_1=2.72$. Then, by (17),

$$U=45-30\log_e\frac{r}{r_1},$$

and by (45B),

$$U_\Delta=0.75[(r/r_1)^2-1] -[2.58+0.30(r/r_1)^2]\log_e(r/r_1).$$

When these equations are evaluated, the results are as shown in Table II.

TABLE II

$\frac{r}{r_1}$	U In Degrees	U_Δ In Degrees
1.00	45	0.00
1.40	35	−0.34
1.95	25	−0.38
2.72	15	0.00

It is evident from these results that U_Δ, and consequently $u_\Delta=U_\Delta\cos\psi$, may be neglected in most practical cases.

When $u=U\cos\psi$ in accordance with (41), equation (42) may be solved directly by a series, the form of which is suggested by the nature of the terms u_Δ, $u_{\Delta\Delta}$, etc. By (42) and (43),

$$\frac{\partial^2u}{\partial r^2}+\frac{1}{r}\frac{\partial u}{\partial r}=-\left(\frac{4\pi}{\lambda}\right)^2u.$$

The series solution is

$$u=\sum_0^\infty\left(\frac{r}{r_1}\right)^{2n}\left(a_n+b_n\log\frac{r}{r_1}\right),\qquad(46)$$

where n has only the values 0 and positive integers. The coefficients a_0 and b_0 may be considered to be the arbitrary constants. Substituting (46) into the differential equation, there results

$$4\sum_0^\infty\frac{r^{2(n-1)}}{r_1^{2n}}\left[(n^2a_n+nb_n)+n^2b_n\log\frac{r}{r_1}\right] =\left(\frac{4\pi}{\lambda}\right)^2\sum_0^\infty\frac{r^{2n}}{r_1^{2n}}\left(a_n+b_n\log\frac{r}{r_1}\right).$$

To satisfy this for any value of r, the coefficient of r^{2k} must equal 0, and also, independently, the coefficient of $r^{2k}\log(r/r_1)$, giving the relationships

$$(k+1)^2\,a_{k+1}+(k+1)b_{k+1}=\left(2\pi\frac{r_1}{\lambda}\right)^2a_k,$$

$$(k+1)^2\,b_{k+1}=\left(2\pi\frac{r_1}{\lambda}\right)^2b_k.$$

From these, the various coefficients can be found in terms of the arbitrary constants a_0 and b_0, which latter are then found by the fact that $u=u_1$ at $r=r_1$, and $u=u_2$ at $r=r_2$. As the factor $(2\pi r_1/\lambda)^2$ appears again, the coefficients evidently diminish rapidly as n (or k) increases. If only the terms for $n=0$ and $n=1$ are retained, computation of the coefficients should yield a result very nearly equal to the sum of (17) and (45B). As different approaches were used in the two solutions, there will be a very slight difference in the results when the higher-order terms are neglected.

For heat generated in the dielectric, the first approximation to the temperature distribution is given by (33B), which is of the same form as (17) but with a term $[\log(r/r_1)]^2$ added. When the process of determination of U_Δ is carried through, the general form of the expression for U_Δ is the same as (45A) with an added term

$$\left(\frac{r}{r_1}\log\frac{r}{r_1}\right)^2.$$

The coefficients of the terms are different from those given in (45B), but the net result is the same: that axial heat flow in the dielectric can be neglected for practical purposes in most cable problems.

The assumption that the temperature of the conductors is a function of x only and not of r could be justified by reasoning, similar to that above for neglecting axial flow in the dielectric, on the basis of the shortness of the radial dimensions of the conductors compared to a wavelength.

Discussion of

Exact Design and Analysis of Double- and Triple-Tuned Band-Pass Amplifiers*

By MILTON DISHAL

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VERNON D. LANDON:[1] Dishal has written an excellent summarizing paper on band-pass amplifier design. I believe it is slightly misleading, however, as to the value of Q required for operation of triple-tuned circuits.

On page 366, in speaking of a band-pass filter utilizing three tuned circuits, Dishal says:

"To obtain a flat-topped response with three peaks of equal amplitude in the pass band, all the loading must be removed from the middle tuned circuit. . . . Otherwise, as will be shown later, the outer two peaks of the response will be lower in amplitude than the middle peak."

In the next paragraph he elaborates: "To approach the ideal triple-tuned response curve, the Q of the middle tuned circuit must be of the order of 10 times (or more) the Q of the input and output circuits."

The experimental facts are somewhat at variance to the above, as will be explained. Given three tuned circuits with $Q_1 = Q_3$ and with $Q_2 = 10Q_1$, the circuits may be coupled to obtain the ideal triple-tuned response curve to which Dishal refers. If, for economy, or other reasons, the value of Q_2 must be reduced to only 3 or 4 Q_1, the outside peaks will have lower amplitude than the middle peaks (as he states), providing no other circuits constants are changed. However, if at this point Q_1 is reduced somewhat and Q_3 is increased (or the reverse), the equality of the three peaks may be restored. This fact is rather important, as it permits the use of three coupled circuits without having to meet quite as stiff a Q requirement as that proposed by Dishal.

The smallest value of Q that may be employed in the circuit having the highest Q may be found by making use of Dishal's mathematics. In (28), if the coefficients of F^4 and F^2 are set equal to zero, we have the condition for "maximal flatness";[2] that is to say, the flattest curve without multiple peaks. This gives the two equations:

$$2K^2 - (n_1^2 + n_2^2 + n_3^2) = 0 \tag{1}$$

$$K^4 - K^2(n_1^2 + n_3^2 - n_2(n_1 + n_3)) + n_1^2 n_2^2 + n_2^2 n_3^2 + n_3^2 n_1^2 = 0 \tag{2}$$

where K = the coupling coefficient, and n_1, n_2, n_3 = the inverse of the Q's of the three circuits.

From (1) it appears that, if one of the n's is increased, another must be decreased. Then the largest value, n_0, required for the smallest n, will occur when the two smaller n's have the same value.

Assuming the two smaller n's are n_2 and n_3, we have $n_0 = n_2 = n_3$, and

$$2K^2 - n_1^2 - 2n_0^2 = 0 \tag{3}$$

$$K^4 - K^2(n_1^2 - n_0 n_1) + 2n_1^2 n_0^2 + n_0^4 = 0. \tag{4}$$

Of the three unknowns, K, n_1, and n_0, any one may be assumed fixed, and the other two may be solved for in terms of that one. Cut-and-try methods yield the following solution, which may be checked by substitution:

$$n_0 = 0.236 n_1$$
$$K = 0.745 n_1.$$

In Dishal's (28), the last term of the polynomial under the radical is

$$\left[K^2\left(\frac{n_1 + n_3}{2}\right) + n_1 n_2 n_3\right]^2$$

and is equal to

$$\left(\frac{f_b}{f_0}\right)^6$$

where f_b = the bandwidth at 70 percent (for the condition of maximal flatness), and f_0 = the resonant frequency. Then

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* The full paper appeared in *Electrical Communication* v. 24, pp. 349–373; September, 1947, as a reprint from *Proceedings of the I. R. E.*, v. 35, pp. 606–626; June, 1947.

[1] Radio Corporation of America, RCA Laboratories, Princeton, N. J.

[2] V. D. Landon, "Cascade Amplifiers with Maximal Flatness," *RCA Review*, Pt. I, pp. 347–363; January, 1941: Pt. II, pp. 481–498; April, 1941.

$$\frac{f_b}{f_0}=\left(K^2\frac{n_1+n_0}{2}+n_1n_0^2\right)^{1/3}$$
$$=0.737n_1.$$

Now, for Dishal's assumed conditions of

$$n_1=n_3$$
$$n_2=0,$$

we find $K=n_1=f_b/f_0$ for the maximally flat condition. Dishal assumes that

$$n_2=\frac{1}{10}\frac{f_b}{f_0}$$

is required.

The present discussion indicates that, when $n_2=n_3=n_0$,

$$n_0=0.236n_1$$
$$=\frac{0.236}{0.737}\frac{f_b}{f_0}$$
$$=0.32\frac{f_b}{f_0}$$

is sufficiently small. In other words, the triple-tuned circuit is operable if tuned circuits are available having a Q as high as about

$$3\frac{f_0}{f_b}.$$

MILTON DISHAL: I would like to take this opportunity to thank Landon for pointing out the fact that it is possible to obtain a triple-tuned response curve having three peaks of equal amplitude and two valleys of equal amplitude, even though the Q of the middle resonant circuit is not infinite. This fact is practically of great importance and I think it is safe to say that when triple-tuned band-pass circuits are used, and a symmetrical band pass is desired when the circuits are correctly resonated, the "Q distribution" pointed out by Landon, i.e., $Q_2=Q_3=AQ_1$, should be used rather than the Q distribution mentioned in my paper of $Q_1=Q_3$ and $Q_2=\infty$. (A is a number whose value depends on the type of response desired.)

In discussing this matter, I think it is important to separate clearly the two types of useful responses which can be obtained, in the following manner: (a) the type of response having n maxima of equal amplitude and $(n-1)$ minima of equal amplitude within the pass band where n is the number of resonant circuits used; and (b) the type of response having a single maximum which occurs at the middle of the pass band.

Response Type (a)

It should be realized that my paper considered this type of response only. The main reason why I was led to consider the Q distribution $Q_1=Q_3$ and $Q_2=\infty$ was that this seemed to be the only distribution which would allow exact design equations to be obtained which were not hopelessly complicated. Unfortunately, this still seems to be the case, and it should be realized that Landon's discussion gives no solution for this type of response. Thus, insofar as the response having peaks and valleys within the pass band is concerned, we have only the qualitative fact that this response can be obtained without the necessity for having $Q_2=\infty$.

However, insofar as practical design is concerned (where exact final values must be experimentally determined), Landon's equations can be used to obtain the transitional shape condition, and the coefficient of coupling can then be increased very slightly to produce a multiple-peaked response. (It will also be necessary to make $Q_2=Q_3$ more than $4.24Q_1$; the greater the peak-to-valley ratio desired, the greater will be the required ratio of $Q_2=Q_3$ to Q_1.)

It may be pointed out here that the following procedure may possibly allow an exact solution to be obtained for the multiple-peaked response, for conditions other than my assumed conditions of $Q_1=Q_3$ and $Q_2=\infty$. As pointed out by Landon, the transitional-shape condition or condition of maximal flatness is obtained when, in my (28), the coefficients of F^4 and F^2 equal zero and when the constant term equals $(\Delta f_{3db}/f_0)^6$. When the multiple-peaked response is desired, the coefficients, rather than equaling zero, must equal some specific value. These specific values can be obtained by substituting, in my (28a), the required values of K and n as obtained from (34), (35), and (36). We can then set the more general coefficients given in (28) equal to the above values obtained through the medium of (28a). It is possible that a usable solution may then be obtained from the three resulting simultaneous equations. Thus, to obtain a certain percentage bandwidth between outside peaks of $(\Delta f_p/f_0)$ with a certain peak-to-valley ratio defined by γ of (35) we find that the coefficient of F^4 must equal $2(\Delta f_p/f_0)^2$; the coefficient of F^2 must equal $1(\Delta f_p/f_0)^4$; and the constant term must equal $[\gamma(1+\gamma^2)(\Delta f_p/f_0)^3]^2$. Thus, using Landon's Q distribution, it will be necessary to solve the three simultaneous equations below in order to find the required ratio of

$$\frac{K}{(\Delta f_0/f_0)}, \quad \frac{n_1}{(\Delta f_0/f_0)}, \quad \text{and} \quad \frac{n_0}{(\Delta f_p/f_0)}$$

to obtain a resulting peak-to-valley ratio whose value is determined by γ.

$$\begin{cases} 2K^2-(n_1^2+2n_0^2)=2\left(\dfrac{\Delta f_p}{f_0}\right)^2 \\ K^4-K^2n_1(n_1-n_0)+n_0^2(n_0^2+2n_1^2)=1\left(\dfrac{\Delta f_p}{f_0}\right)^4 \\ K^2\dfrac{n_1+n_0}{2}+n_1n_0^2=\gamma(1+\gamma^2)\left(\dfrac{\Delta f_p}{f_0}\right)^3. \end{cases}$$

Response Type (b)

In his discussion, Landon has given the exact solution for the constants required to give the limiting case of this single-peaked type of response. (It should be noted that the required conditions for maximal flatness could also be obtained by equating simultaneously to zero the two parts of (30).)

As mentioned previously, Landon's solution is of great practical importance because of the relatively small value of Q required in the two high-Q circuits. When identical circuits are cascaded, the required Q_1 for a given $(\Delta f_{3db}/f_0)$ will be even smaller than $0.737(\Delta f_{3db}/f_0)$, and, therefore, the high-Q circuits $(Q_2=Q_3)$ whose Q must equal $4.24Q_1$ will require a necessary Q even less than $3(f_0/\Delta f_{3db})$. For example, for five cascaded triple-tuned circuits, the required Q_1 is

$$Q_1=0.54(f_0/\Delta f_{3db})$$

and, therefore, the required $Q_2=Q_3$ is only $Q_{2,3}=2.3(f_0/\Delta f_{3db})$.

Since the equations are quite simple, I think it would be helpful to tabulate the equations that enable the complete and exact design to be accomplished for cascaded single-, double-, and triple-tuned band-pass circuits using the maximally flat type of response.

N Cascaded Triple-Tuned Circuits

$$Q_2=Q_3=4.24Q_1$$

$$K=\frac{0.745}{Q_1}$$

$$\frac{Q_1}{f_0/\Delta f_{3db}}=0.737[2^{1/N}-1]^{1/6}$$

$$\frac{V_0}{V}=\left[(2^{1/N}-1)\left(\frac{\Delta f}{\Delta f_{3db}}\right)^6+1\right]^{N/2}$$

or

$$\frac{\Delta f}{\Delta f_{3db}}=\frac{1}{[2^{1/N}-1]^{1/6}}\left[\left(\frac{V_0}{V}\right)^{2/N}-1\right]^{1/6}$$

$$\frac{\text{Gain}_{\text{(per stage)}}}{g_m/4\pi\Delta f_{3db}\sqrt{C_1C_3}}=1.03[2^{1/N}-1]^{1/6}$$

$$\tan\theta_{\text{ (per stage)}} = \frac{-\left(\dfrac{\pm\Delta f}{\Delta f_{3db}}\right)\left[\left(\dfrac{\Delta f}{\Delta f_{3db}}\right)^2-\dfrac{2.01}{[2^{1/N}-1]^{1/3}}\right]}{\left[\dfrac{1.93}{[2^{1/N}-1]^{1/6}}\left(\dfrac{\Delta f}{\Delta f_{3db}}\right)^2-\dfrac{1}{[2^{1/N}-1]^{1/2}}\right]}.$$

N Cascaded Double-Tuned Circuits

$$Q_1=Q_2$$

$$K=\frac{1}{Q}$$

$$\frac{Q}{f_0/\Delta f_{3db}}=1.414[2^{1/N}-1]^{1/4}$$

$$\frac{V_0}{V}=\left[(2^{1/N}-1)\left(\frac{\Delta f}{\Delta f_{3db}}\right)^4+1\right]^{N/2}$$

or

$$\frac{\Delta f}{\Delta f_{3db}}=\frac{1}{[2^{1/N}-1]^{1/4}}\left[\left(\frac{V_0}{V}\right)^{2/N}-1\right]^{1/4}$$

$$\frac{\text{Gain}_{\text{ (per stage)}}}{g_m/4\pi\Delta f_{3db}\sqrt{C_1C_2}}=1.414[2^{1/N}-1]^{1/4}$$

$$\tan\theta_{\text{ (per stage)}}=\frac{\mp\left[\left(\dfrac{\Delta f}{\Delta f_{3db}}\right)^2-\dfrac{1}{[2^{1/N}-1]^{1/2}}\right]}{\pm\left[\dfrac{1.414}{[2^{1/N}-1]^{1/4}}\left(\pm\dfrac{\Delta f}{\Delta f_{3db}}\right)\right]}.$$

N Cascaded Single-Tuned Circuits

$$\frac{Q}{f_0/\Delta f_{3db}}=[2^{1/N}-1]^{1/2}$$

$$\frac{V_0}{V}=\left[(2^{1/N}-1)\left(\frac{\Delta f}{\Delta f_{3db}}\right)^2+1\right]^{N/2}$$

or

$$\frac{\Delta f}{\Delta f_{3db}}=\frac{1}{[2^{1/N}-1]^{1/2}}\left[\left(\frac{V}{V}\right)^{2/N}-1\right]^{1/2}$$

$$\frac{\text{Gain}_{\text{ (per stage)}}}{g_m/2\pi\Delta f_{3db}C}=[2^{1/N}-1]^{1/2}$$

$$\tan\theta_{\text{ per stage}}=\left(\pm\frac{\Delta f}{\Delta f_{3db}}\right)[2^{1/N}-1]^{1/2}.$$

Contributors to This Issue

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Christian W. Hirsch was born in 1905. He was graduated as an electrical and telephone engineer from the Technical University of Norway.

After two years with Siemens & Halske, he started work with Standard Telefon og Kabelfabrik A/S in 1930 in the telephone switching installation department. From 1931 to 1932, he was engaged in repeater installation work, and from 1932 to 1936 in the telephone and power cable test room in the factory. After two years in the cable sales department, he was transferred to the cable engineering and installation department. Since 1946, he has been in charge of manufacturing engineering for the telephone and cable factories.

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Erik H. L. Julsrud was born in Oslo in 1902. After graduation from the Norwegian Institute of Technology in Trondheim, he spent three years as a private assistant to the professor of physics at the institute.

He joined the International Standard Electric Corporation laboratories in London early in 1930. Later he was transferred to Les Laboratoires Standard in Paris and from there to Standard Electric A/S, Oslo.

Mr. Julsrud left the I.T.&T. System in 1932 and, after a year as a consulting engineer, joined the technical staff of the Norwegian broadcasting system. There he has been in charge of the development of audio-frequency facilities and was responsible for the planning and construction of the audio-frequency equipment in the new broadcasting house.

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Gerard J. Lehmann was born in Paris, France, on April 6, 1909. He received an engineering degree from Ecole Centrale in 1931.

On leaving school, he entered the employ of Sadir, becoming technical director in 1939.

In 1940, after leaving the French Army, he became a member of the Lyon laboratory staff of Le Matériel Téléphonique. In 1943, he joined Federal Telephone and Radio Laboratories in New York.

Mr. Lehmann returned to France in 1945 as a consulting engineer for Laboratoire Central de Télécommunications. In addition to his research work, he has been teaching at Ecole Centrale and in 1942 was appointed professor of direction finding and radio navigation at Ecole Supérieure d'Electricité.

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William W. Macalpine was born in Scotland, September 2, 1900. In 1922 he received the B.S. degree in electrical engineering at Carnegie Institute of Technology. He received the M.A. in 1924 and Ph.D. in physics in 1930 from Columbia University, where for three years he was assistant in the Ernest Kempton Adams Precision Laboratory of Optics and Electricity.

Except for a few months with Wired Radio in 1931, he has been continuously associated with the I.T.&T. System since 1929. He served on the engineering staffs of International Communications Laboratories, Federal Telegraph Company, Federal Telephone and Radio Corporation, and, since 1947, in the radio and radar components division of Federal Telecommunication Laboratories. Much of his work has been on communication and marine radio receivers, marine radio direction finders, and low-frequency radio range equipment.

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B. Secker was born in Dewsbury, Yorkshire, in 1910 and obtained an open mathematical exhibition to Magdalene College, Cambridge, in 1929. He obtained a First Class in Part 1 of the Mathematical Tripos in 1930 and A Senior Optime in Part 2 of the Tripos in 1932, obtaining his B.A. in 1932 and M.A. in 1940.

He joined Standard Telephones and Cables, Limited, in 1936 as a development engineer in the transmission division, and has since then been engaged in theoretical investigations in connection with the design of testing apparatus. He became an Associate of the Institute of Physics in 1940.

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André Emile Violet was born in Paris, August 10th, 1913.

He studied at the Lycée Charlemagne, Paris, and later at the faculté des Sciences where, in 1934, he graduated with a degree in Physical Science. He obtained the diploma of advanced study in physics and chemistry in 1935 on a physiochemical subject.

While preparing his fellowship in physics and chemistry, which he received in 1938, he was employed as a replacement teacher.

Mobilized as a military chemist, he experimented on methods of protection against toxic gases. Demobilized in August, 1940, he was appointed professor at the Bayonne lycée.

He left the teaching profession in 1941 to accept employment by Le Matériel Téléphonique as a radio laboratory engineer. First specializing in radio-electric measuring apparatus, he later worked on television, frequency-modulation receivers, mine detectors, and aviation radio-electrical apparatus.

He has taken out various patents, the most important of which relate to power regulating devices and mine detectors.

• • •

T. A. G. **Weider** was born in Trondheim, Norway, in 1903. He was graduated from the Norwegian Institute of Technology in Trondheim in 1927.

After graduation, he was employed by Standard Electric A/S, Oslo, where he served with the telephone division. After two years, he was transferred to the company's office in Copenhagen, later Standard Electric A/S, Copenhagen. In 1934, he returned to Oslo and is now in charge of the commercial radio division.

Mr. Weider is a member of the Norwegian Society of Civil Engineers and of the Norwegian Electrotechnical Society.

• • •

C. **Gordon White** was born in Dublin, Eire. He received the Diploma in Engineering (A.R.C.Sc.I.) of the Royal College of Science for Ireland in 1919. He continued at the college in tutorial duties until 1923. In 1920, he obtained the B.Sc. (Eng.) degree of London University with Honours.

In 1924, Mr. White joined the Western Electric Company in London. After a year in the telephone circuit department at North Woolwich he transferred to the circuit laboratory. In 1930, he joined the remote-control department of Standard Telephones and Cables, Limited, where he has been engaged in technical sales, design, and planning of supervisory control and telemetering systems.

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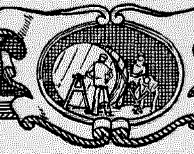

ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies

COPENHAGEN BROADCASTING HOUSE
ZURICH AUTOMATIC TANDEM TOLL EXCHANGE
EMERGENCY OPERATION OF SOS–12 CARRIER SYSTEMS
NICKEL–IRON ALLOY DUST CORES
VERY-HIGH-FREQUENCY SINGLE-CHANNEL RECEIVER
HIGH GAIN WITH DISCONE ANTENNAS
DEMONSTRATION OF MACDONALD'S EQUIVALENCE THEOREM
CIRCULAR WAVE GUIDES WITH TWO COAXIAL DIELECTRICS
HIGH-RATIO MULTIVIBRATOR FREQUENCY DIVIDER
PROPERTIES OF BLUE TITANIUM DIOXIDE
TRANSMISSION RATE, BANDWIDTH, SIGNAL-NOISE RATIO
NEW DEVELOPMENTS IN MARINE RADIO DIRECTION FINDERS
H. T. KOHLHAAS, EDITOR, RETIRES

JUNE, 1948

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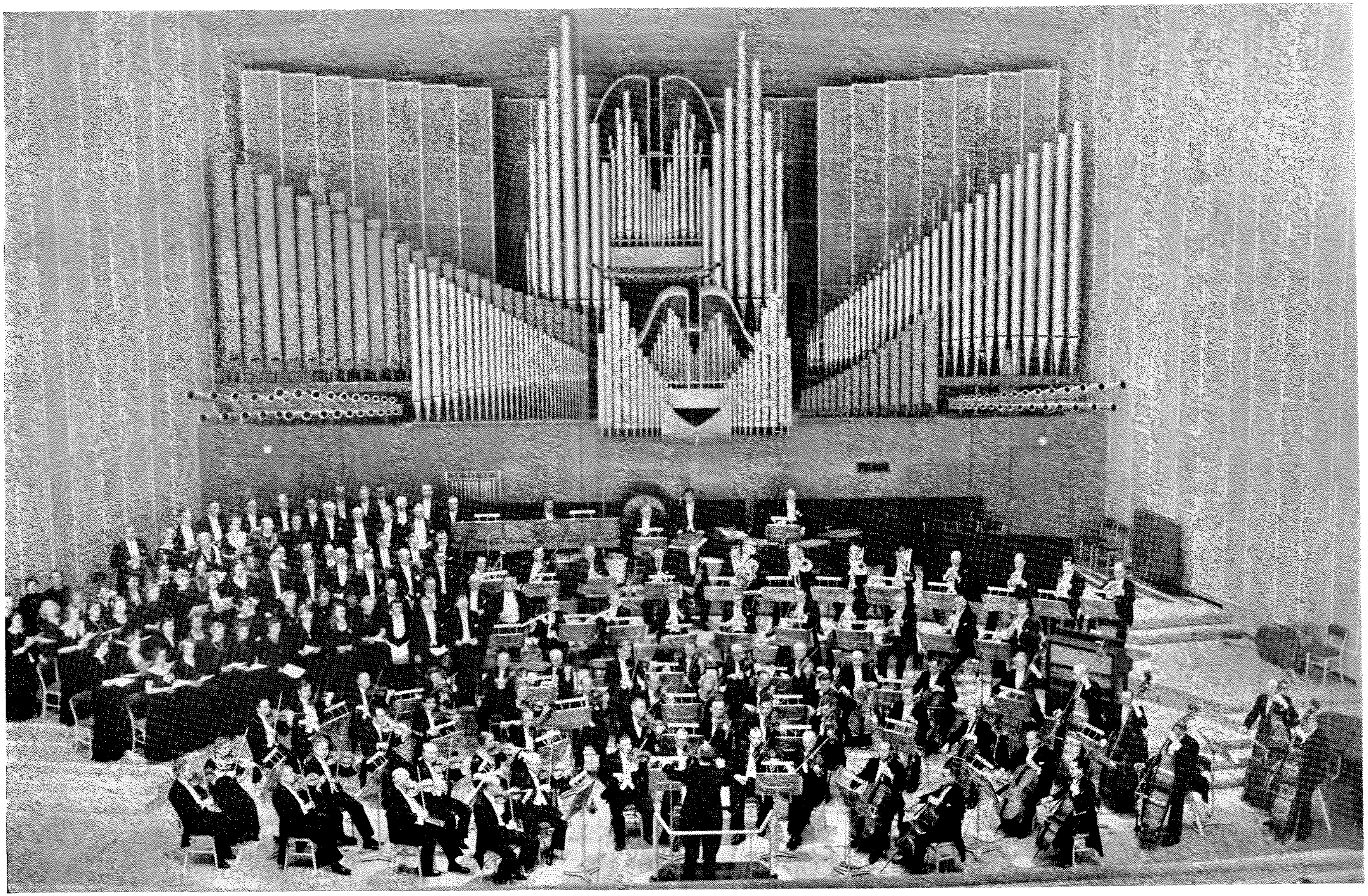

The Great Concert Studio of Copenhagen Broadcasting House. The stage will accommodate an orchestra and choir of 100 persons each.

Copenhagen Broadcasting House

By FREDERIK HEEGAARD
Post and Telegraph Administration, Copenhagen, Denmark,

and EIGIL COHRT
Standard Electric A/S, Copenhagen, Denmark

COPENHAGEN BROADCASTING HOUSE was built and put in service during the war. In planning the buildings, careful attention was given to obtaining adequate sound insulation among the studios. This was obtained by building each studio as a separate structure with its own foundation. In addition, Broadcasting House is near the center of town and noises from the streets surrounding it must not penetrate to the studios. These street noises are reduced by screening the studios with the office and administration buildings, the concert hall, and by other structures. This arrangement is evident from Fig. 1.

1. Studios

The studio block is divided through the middle by a wide hall with bays, arranged as a lobby for the artists. From the lobby, a studio is entered by passing through at least two soundproof doors connected by an acoustically heavily damped corridor, which acts as a sound-trap. Associated with each studio is a control room intended for supervision and control of broadcasts and rehearsals in the studio; access to this room is also obtained from the sound-trap. Large triple glass windows provide visibility between the studios and their control rooms without sound leakage. In several cases, a second studio control room is common to two studios, thereby making it possible to operate them as one. Studios 2 and 3 are an example of this. Table I gives data on studios 2 to 8.

In addition, there are a concert hall and three small lecture studios. The former, which is shown in the frontispiece, will accommodate 100 musicians, a choir of 100, and about 1200 auditors. Its volume is 12,000 cubic meters (423,700 cubic feet). The auditors may enter directly from the street. The lecture studios are in the administration block with direct access for the lecturers from the main hall. To obtain the necessary sound insulation, these three studios are provided with floating floors, walls, and ceilings. Two echo rooms of different characteristics are in the studio block.

The technical department is located centrally above the studio hall; it consists of a large amplifier room with the amplifiers mounted in two rows of steel cabinets, a room containing the switching system for connections to the studios, three main control rooms, four sound-recording rooms, and a couple of offices for the operating management. This central position of the technical installations is convenient in view of the fairly complicated cabling between the studios, studio control, main control, amplifier, recording, and relay rooms.

2. Program Circuits

All program circuits have been laid as single-pair paper-insulated cables with a special screen insulated from the lead cover. To decrease cross talk, the pairs have been twisted with two

TABLE I

STUDIO CHARACTERISTICS

Studio	Service	Volume		Reverberation Time in Seconds at 800 Cycles
		Cubic Meters	Cubic Feet	
2	Orchestra, Medium Size	2480	87,669	1.1
3	Orchestra, Small	1450	51,200	Variable
4	Soloist, Chamber Music	475	16,782	0.7
5	Discussions, Small Cabaret with Audience	730	25,776	0.8
6	Soloist, Plays	414	14,618	0.5
7	Plays	1050	47,076	0.5
8	Speech in Plays	47	1,660	0.5

different pitches, one for cables carrying high signal levels and the other for low-level cables. A view of the amplifier room is shown in Fig. 2.

The amplifier equipment may be divided into the following main groups: (A) program sources, (B) switching and signaling systems, (C) program channels, (D) output patching panel, (E) control apparatus, (F) sound recording system, and (G) talk-back equipment.

A program source may consist of a studio or a combination of studios with their microphones and amplifier system, a telephone line transmitting an outside program, or a playback apparatus for recorded programs. In each studio, a number of microphones may be used to fit the needs of the particular type of program. Each microphone is connected to an *A* amplifier, all of which are located in the amplifier room. The *A* amplifiers are one-stage push-pull units, three of which are mounted on a single panel and share a common power supply as may be seen in Fig. 3. The signal from the *A* amplifier then passes through a bridged-T mixing pad mounted on a desk, shown in Fig. 4, in the studio control room. Next in the circuit is a *B* amplifier. This is a three-stage resistance-coupled amplifier with negative feedback, containing its own power supply. Then follows the main regulating fader, also mounted on the studio control desk, which is connected through

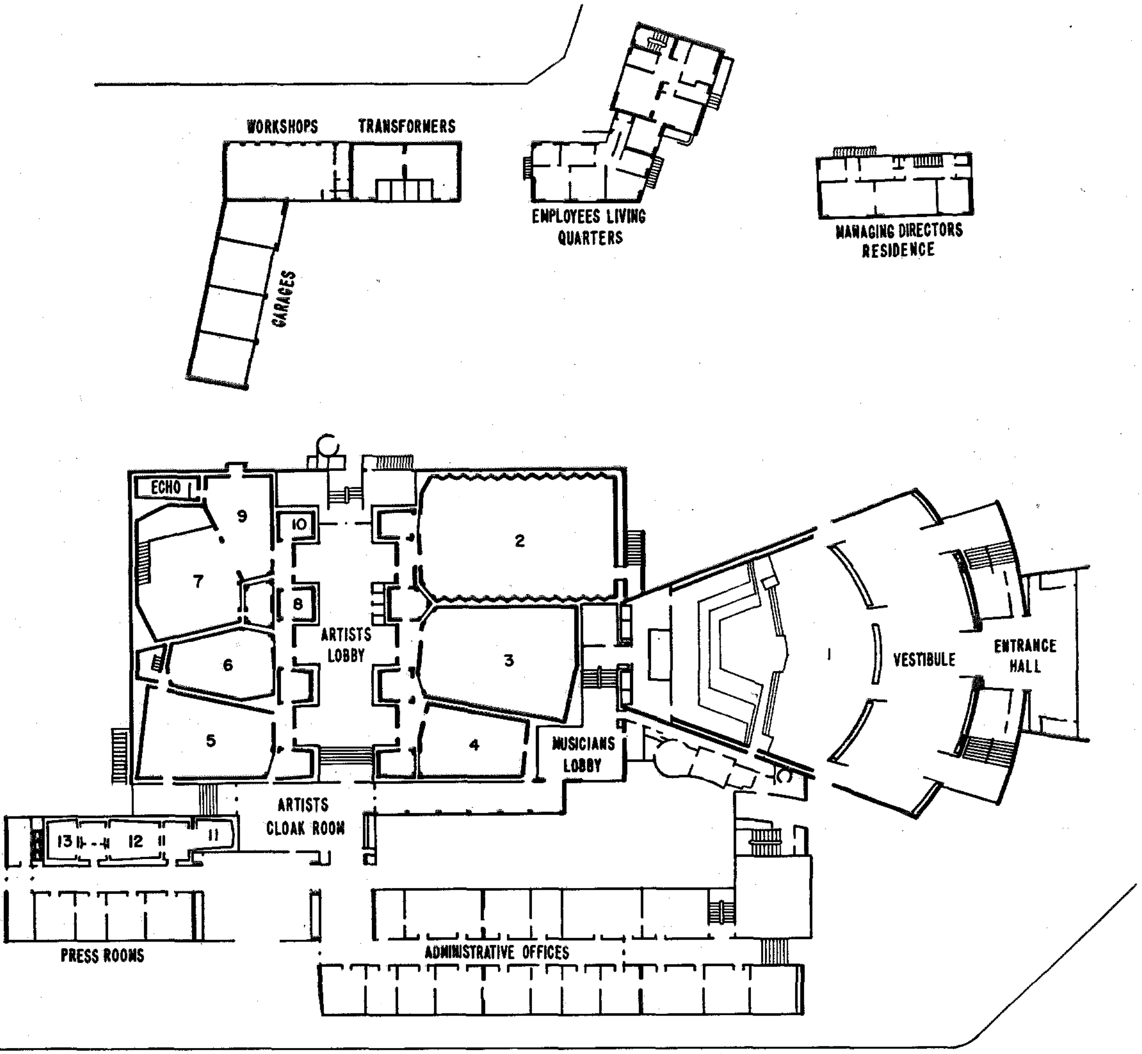

Fig. 1—Plan view of Copenhagen Broadcasting House. The central studio building is screened extensively from city noises by the various administrative and other buildings surrounding it.

a transformer to an appropriate switching system.

The lines carrying outside programs to Broadcasting House arrive in two cables, which also contain the outgoing lines to the transmitters and to the Copenhagen repeater station. The outside-broadcast lines are connected through line transformers, variable equalizers, attenuators, and, if required, *A* amplifiers, before going to the switching points.

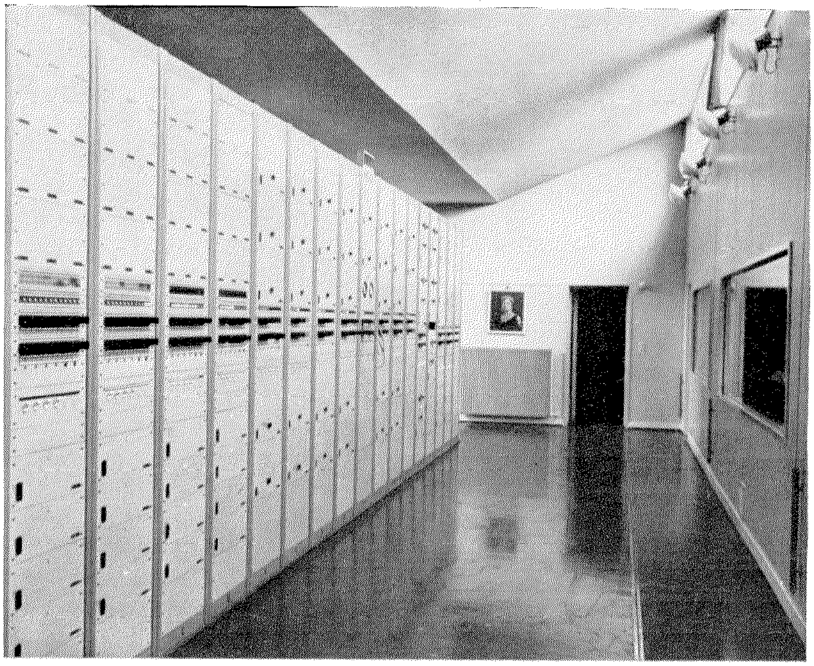
Fig. 2—Part of the amplifier room.

In principle, the playback apparatus in the recording rooms and elsewhere are connected to the switching system in the same way as described for the studios. In this instance, *A* amplifiers have been omitted, the level being higher. They are provided with the necessary equalizing networks and scratch filters.

The programs pass from the studio controls, through the switching system, into one of three program channels, each having four inputs. Each channel input is provided with a listening key and mixer, while each channel has also a main attenuator. In addition, there are three channels of a simpler type having one input only and no attenuators. These three so-called "common channels" are used for rehearsals and for uncomplicated sound recordings requiring no special arrangements.

To stage, supervise, and control complicated broadcasts involving several studios, an arrangement called the "dramatic control" has been developed. It consists of a special control desk with six input channels and one output with the necessary controls. It may be connected between the studio control and the main control by the automatic switching system, and provides for full control of the broadcast. The dramatic control can handle programs from all studios, but is arranged to provide visual supervision of studios 6, 7, 8, and 9. This dramatic control also permits the introduction of variable equalizers to give special effects to certain studios or echoes. A studio amplifier similar to the *B* amplifier compensates for the losses introduced in the mixer.

3. *Switching*

All switching between program sources and channel or dramatic inputs is done automatically. The switching of microphone leads, volume controls, and signaling system is done in one operation, reducing substantially the possibilities for mistakes.

The switching system must have a very high degree of reliability, and at the same time must

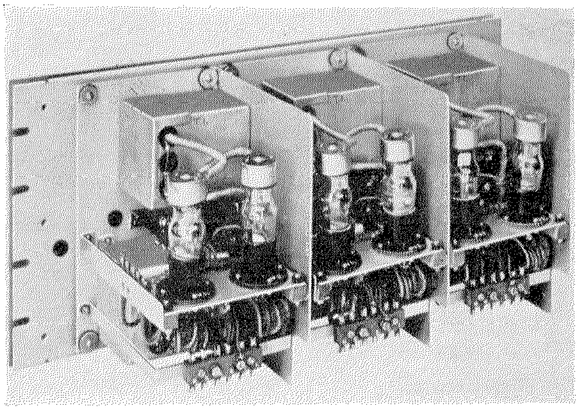
Fig. 3—A group of three *A* amplifiers mounted on a single panel. These amplifiers are connected to the microphones to provide both gain and isolation.

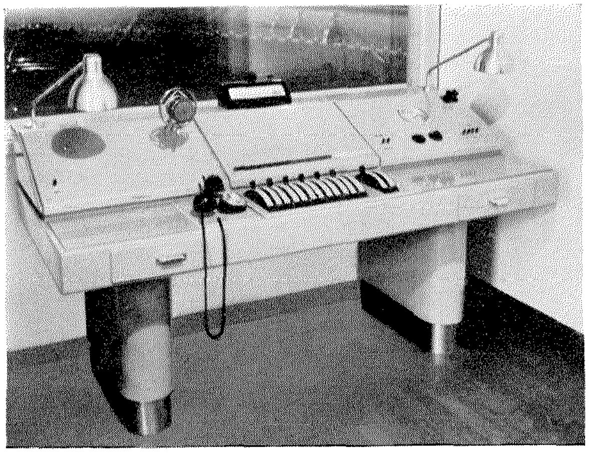

Fig. 4—Studio control desk.

be very flexible and simple in operation. Because of the amplification employed, switching must be entirely free from induced, capacitive, and microphonic noises. There are 21 channel inputs to which a maximum of 100 program sources may be connected. An entirely satisfactory system has been built with the same kind of apparatus and switches used in public and private automatic rotary telephone systems of Standard Electric manufacture. All relays are provided with split contact springs and double spherical contacts.

Switching is based on a coordinate system with relays at the intersecting points. The selection of the proper relay is determined by a control circuit that receives key-sent numbers and selects and operates the correct relay. The relay locks electrically and releases the control circuit for another call. To minimize the number of digits to be sent and to safeguard against mistakes, a channel of a particular control desk may be selected only from that desk. A distinction is also made between selection of a program and selection of an echo room; a three-position key in one position will permit only program selection and in another only echo-room selection.

Progress lamps on the control desks indicate when the control circuit is ready to receive the digits and when the selection is finished. Busy lamps indicate the channels in use. As an extra safeguard, each program channel or echo room is provided with a two-digit number-indicator operated from the control circuit during selection. It permits the operator to check immediately that the correct number has been sent and that the appropriate selection has taken place. The indicator remains in position until the connection is released, thus permitting the operator at any time to know which program source is connected to any particular channel. This relay connection is under the control of a locked release key on the main control desk, and the relay may be released by depressing this key momentarily.

Program connections may be set up prior to starting a broadcast by opening the attenuators of the channel, thereby permitting a change from one program to another by further operation of the attenuators only, without waiting for selection of circuits by relays.

In case of faults on lines, attenuators, or connecting relays, a program may be transferred from one channel to another on the same or another control desk without interruption. The new connection is made while the old is still in service, and the program faded from the faulty circuit to the new one.

Routine test circuits have been provided by which all the functions of the control circuits may be checked. Faults that might cause delay

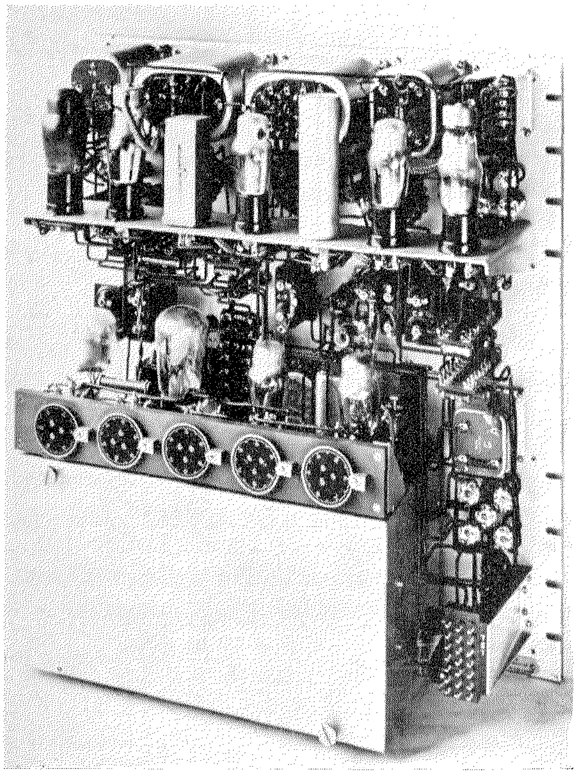

Fig. 5—The two-stage *C* amplifier, which also provides limiting action to prevent overmodulation. Two *C* amplifiers are mounted on a single panel.

in operation may thus be avoided. Two control circuits are provided and are utilized alternatively; thus, if one should be faulty, the second is always ready for use.

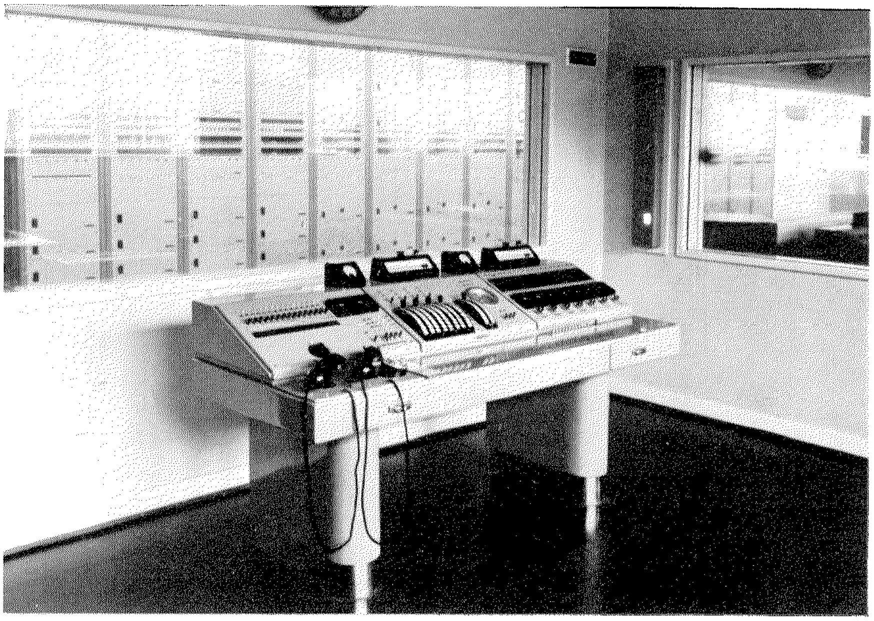

Fig. 6—Main control desk. The amplifier room is visible to the operator through large triple windows.

Particular care is taken to insure the system against cross talk and noise, by special screening and by keeping microphone and signaling leads completely separated throughout the system. No such troubles have been observed during the five years of operation.

A *C* amplifier is connected between the program source and the output switching panel. The *C* amplifier, shown in Fig. 5, consists of two two-stage resistance-coupled amplifiers, connected through a limiter to prevent the output level from rising above the value that corresponds to full modulation of the transmitter.

The program channels are operated from the three main control rooms. These are adjacent to the amplifier room, which may be viewed through large triple glass windows. Each of the three control rooms may operate one of the three program channels as well as all of the common channels. In each control room, all the different operations may be accomplished at the main control desk shown in Fig. 6.

For visual supervision of the program level, each channel output has been connected to a modulation meter with instruments on main and studio control desks. In principle, the modulation meter or tone meter is a rectifier obeying a logarithmic law and having a quick charge and slow discharge characteristic, so that peak voltages may be read. They are light-beam instruments with a very short time constant; being calibrated in nepers, the scale is practically linear. The upper part of the scale, corresponding to modulation of the transmitters above 80 percent, is red so that danger of overmodulation will be immediately apparent. When a studio is connected to a channel input, the instrument in the corresponding studio control room is automatically connected to the modulation meter of the channel in use.

Each control desk includes an output patching panel where, by means of a manual coördinate switchboard, the outputs of the *C* amplifiers associated with the channels are connected to the inputs of *D* amplifiers connected to the outgoing lines.

The *D* amplifiers are two-stage resistance-coupled amplifiers with low output impedance, so that they supply constant voltage to lines of varying impedances. Their outputs are connected to the "program consumers," i.e., the broadcasting transmitters, the Copenhagen repeater station for broadcasts to foreign countries, the recording rooms, and the loudspeaker channels.

The lines used for recording pass from the *D* amplifiers to the four recording studios through special recording amplifiers, which are power amplifiers similar to the loudspeaker amplifiers. Two of the recording studios are each provided with two wax recorders and two cellulose recorders, and have in addition play-back apparatus for cellulose and ordinary commercial records. The wax recorders are combined with play-back apparatus. The other two sound-recording studios are provided with tandem magnetophone equipment to permit continuous magnetic recording and reproduction using plastic tape. A reproducer for commercial sound motion pictures is also available. A small announcer's studio is provided near the recording rooms to be used when recording or playing back programs.

Loudspeakers have been installed in studios, studio control rooms, and main control rooms for supervision of programs and for talk-back from studio control rooms to studios. The input to this system is taken from a separate *D* amplifier connected to each channel output. Feedback in the studios is avoided by attenuating the studio loudspeaker as soon as the attenuator of any microphone in the studio is opened. When talking back from control room to studio, a microphone placed on the control desk is cut in by a key, which also attenuates the control-room loudspeaker output.

A loudspeaker distribution system serves all offices, Council Hall, canteen, etc. The system accommodates four program channels, thus enabling the staff to listen to rehearsals as well as the running program. The loudspeakers require no amplifiers as the program lines are operated at high level from 40-watt amplifiers manufactured at the shops of the Post and Telegraph Administration.

Over the frequency range from 30 to 10,000 cycles per second, the total amplification from microphone to line does not deviate more than ±0.2 neper (1.7 decibels) from the value at 800 cycles. At a total amplification in the system of 9.5 nepers (82.5 decibels), the noise voltage at the output of a *D* amplifier will not exceed 5 millivolts. In the range from 100 to 10,000 cycles, the distortion factor for the complete amplifier chain must not exceed 1 percent and from 30 to 100 cycles it must be less than 4 percent. The distortion factors are, in fact, considerably lower.

4. Signaling

A signaling system provides for the essential cooperation among all factors concerned with a broadcast transmission. The automatic switching system connects all the signaling and program circuits at the same time, thereby assuring readiness of the signaling system when required and avoiding separate operations with their possibilities of error.

The system includes red warning lamps and blue end-of-program lamps together with the necessary operating keys. The red lamps are displayed in the studio proper, on a signal stand in the studio, outside the studio entrance, on the studio control desk, and on the channels on the main control desk. The blue lamps are displayed on the studio signal stand, the studio control desk, and on the channels of the main control desk. Mounted on the wall in the three control rooms and in the sound recording rooms are signal boards on which the red and the blue lamps of each studio are represented.

An instant before a program is to go on the air, the control-desk operator—studio or main—in charge of this broadcast gives a microphone warning signal by flashing the red lights and then operating them steadily. In the case of delay in the studio, the studio manager may operate the microphone blocking key and continue the flashing, thereby indicating that the connection may not yet be made. When ready, he will restore his key and produce a steady red light. At the end of the program, the studio manager will switch on a blue light, the end-of-program signal, and the control desk will open the connection, thereby extinguishing the red light.

Connections among the various points of importance, such as studios and control desks may also be obtained by telephone through a private automatic branch exchange. As it may be of importance to obtain immediate communication with a certain party, the connection will be established even if the called line is busy; a warning tone will indicate that a third party is on the line.

The studios, buildings, and amplifier equipment were designed by the Broadcast Department of the Post and Telegraph Administration and manufactured by Standard Electric A/S, Copenhagen, who also designed and manufactured the switching equipment, desks, loudspeaker cabinets, and similar facilities.

Zurich Automatic Tandem Toll Exchange

By G. KLINGELFUSS

Standard Telephone and Radio S. A., Berne, Switzerland

DURING the past 20 years, very great progress has been made in the automatization of the telephone service in Switzerland. Now, about 95 percent of the 450,000 subscribers have automatic switching service, and within a few years the entire network will be automatic.

After the city and rural networks were converted to automatic operation, most of the fundamental problems of long-distance automatic telephony, such as impulse transmission over long lines and metering of calls by time and distance, were solved. The next advance was to provide automatic service between subscribers of different network groups.

As a temporary solution, long-distance dialing by operators was introduced, whereby the toll operator at the outgoing end dialed directly the number of the wanted subscriber in another city.

The first application of direct automatic toll switching by subscribers in different city areas equipped with rotary-system exchanges was introduced in 1933 between Basle and Zurich. Similar services were provided between Berne and Lausanne with Strowger equipment. The first application of interworking between two different systems was between Lausanne and Geneva with Strowger and rotary equipment, respectively.

The good results obtained on these automatic toll circuits encouraged their extension. A fundamental plan for automatic operation of the Swiss toll network had been prepared by the Swiss administration and was elaborated on jointly by Bell Telephone Manufacturing Company and Standard Telephones and Cables, Limited, giving full consideration to switching, signaling, and transmission requirements.

The Swiss telephone network includes three different automatic systems, Hasler, rotary, and Siemens. It was necessary, therefore, to establish fundamental requirements for their interworking. For this purpose, the Swiss administration prepared a master specification that contained all the requirements essential to reliable interoperation.

These specifications concerned not only signaling, but transmission requirements over long-distance lines, with fixed repeaters and with switched repeaters connected at tandem points. The repeater problems were still to be solved, particularly, switched repeaters with automatic gain control. The Swiss administration adopted for all three systems the principles of switched repeaters developed and recommended by Bell-Standard.

The three manufacturers were requested to submit models of every circuit involved in automatic toll switching. The three models were interconnected over real toll-line circuits of the greatest length met with in Switzerland using both fixed and switched repeaters. For signaling, 50-cycle-per-second alternating current was used.

The Swiss automatic toll network in its final stage is shown in Fig. 1. It is comprised of 10 primary centers and about 40 secondary or end exchanges. Each toll center is equipped with repeaters and associated switching apparatus. The toll end exchanges have direct trunks to the toll centers to which they are assigned.

Three-digit prefixes starting with 0 were assigned to the primary and secondary exchanges (Zurich 051). The third digit determines whether the call is destined to remain in the primary center group (51) or has to be extended to an end exchange assigned to this center (Winterthur 052, Schaffhausen 053).

1. Toll Center Zurich

Zurich is the largest and most important toll center of Switzerland. It handles about 20 to 25 percent of the total telephone traffic and consists of 11 city exchanges of 10,000 lines with 3 satellite exchanges of the 7-A1 and 7-A2 rotary system and 50 rural exchanges of the 7-D type. It may be interesting to note that the first rotary exchange at Zurich-Hottingen was put in service in 1917 and is still in operation.

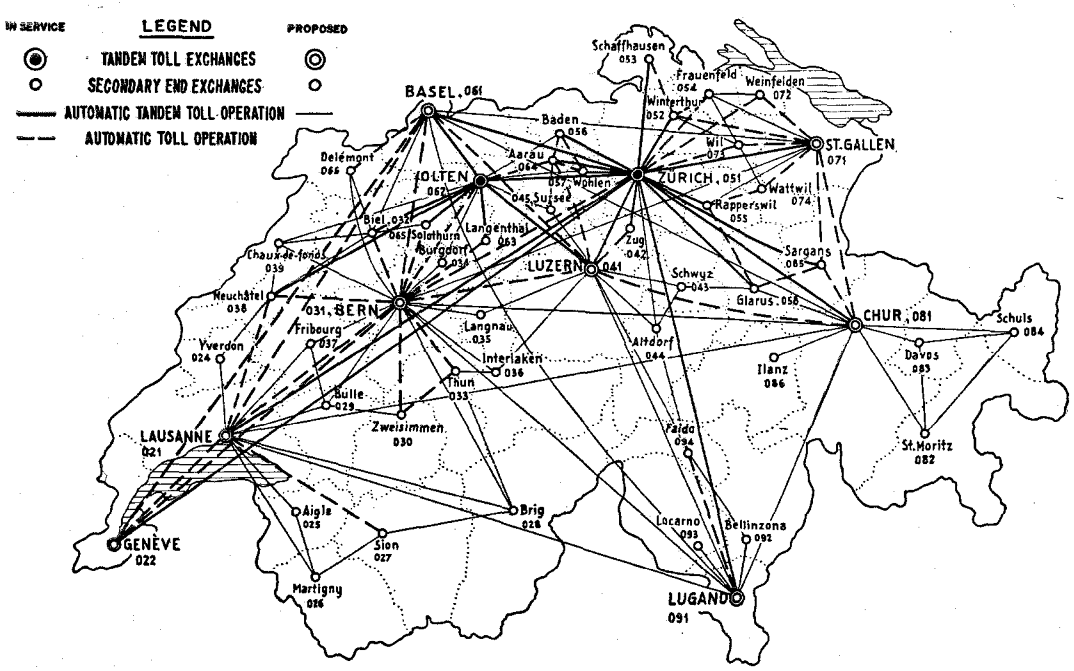

Fig. 1—Automatic toll network of Switzerland.

Zurich also has a very important national and international toll traffic, being the largest commercial and industrial center of the country. Consequently, it is also the most important toll switching point.

The first equipment for the present automatic toll exchange was ordered in 1937. A new building, shown in Fig. 2, was erected for this purpose next to an existing building, in which the manual toll exchange,

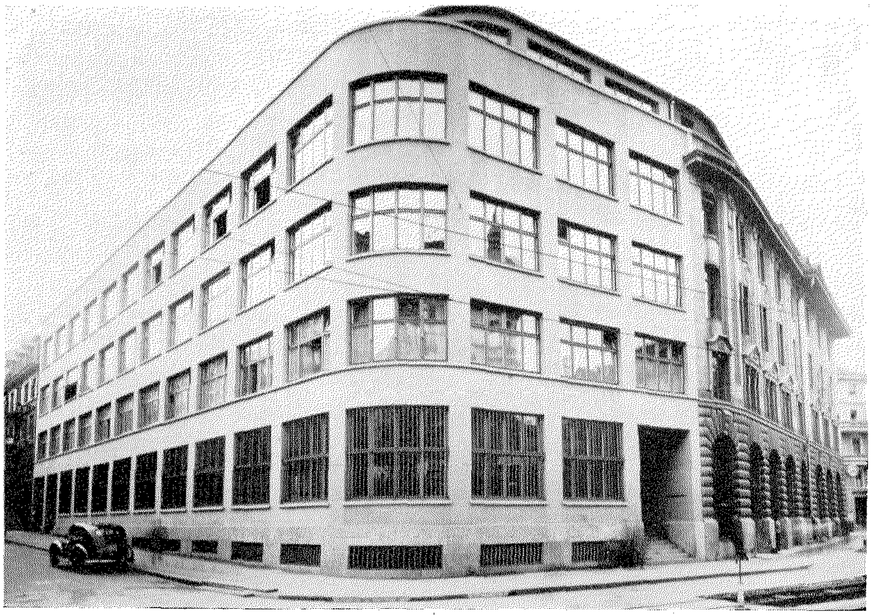

Fig. 2—Building housing the automatic toll exchange and repeater station in Zurich.

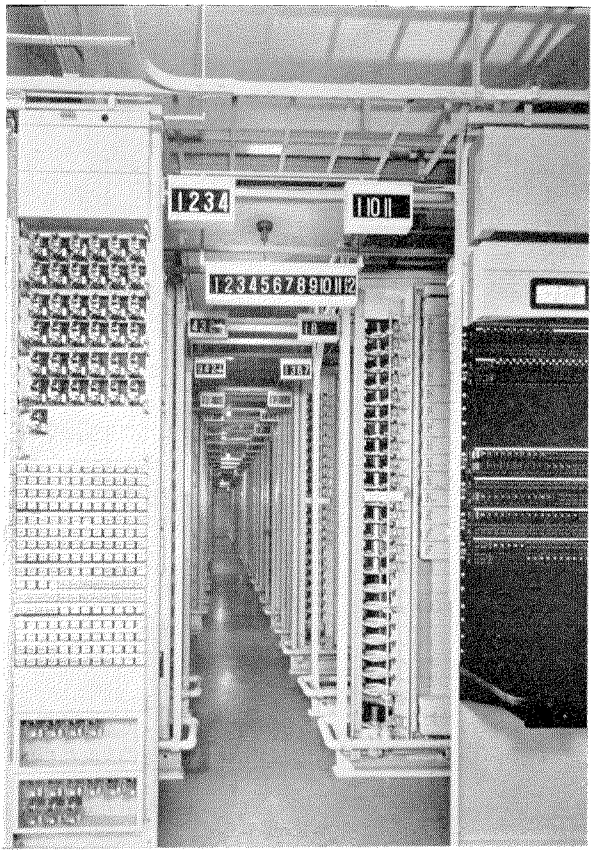

Fig. 3—General view of the automatic toll exchange.

rural main exchange, and repeater station are located.

The automatic toll equipment, shown in Fig. 3, is located on the third floor, and the repeater station is on the floor directly below. This arrangement requires only very short cable runs to the repeaters, which are located in both the automatic toll switching room and in the repeater station.

The first installation of automatic toll equipment, which was placed in service in 1939, consisted only of apparatus for the outgoing terminal toll traffic originated by Zurich city and rural subscribers and of equipment for handling connections established by manual toll operators on incoming national or international toll lines. The equipment for incoming terminal traffic destined for Zurich city and rural subscribers is located in the old building and was also extended and adapted to the administration's master specification.

During the war, several extensions were made and, in 1945, equipment for automatic tandem toll operation was put in service, whereby for the first time automatic switched repeaters were used in the rotary system. Through this development, Zurich became the largest and most important tandem point for automatic toll switching.

A simplified junction diagram of the Zurich automatic toll exchange in its present stage is shown in Fig. 4. The actual equipment occupies a floor space of 550 square meters (5918 square feet), and completely fills the room. Future extensions will have to be located on the floor above.

2. *Toll Traffic Originated by City and Rural Subscribers*

To establish an automatic toll connection, a subscriber first dials a three-digit toll prefix (021 to 099) followed by the number of the wanted subscriber, which may be of two, five, or six digits.

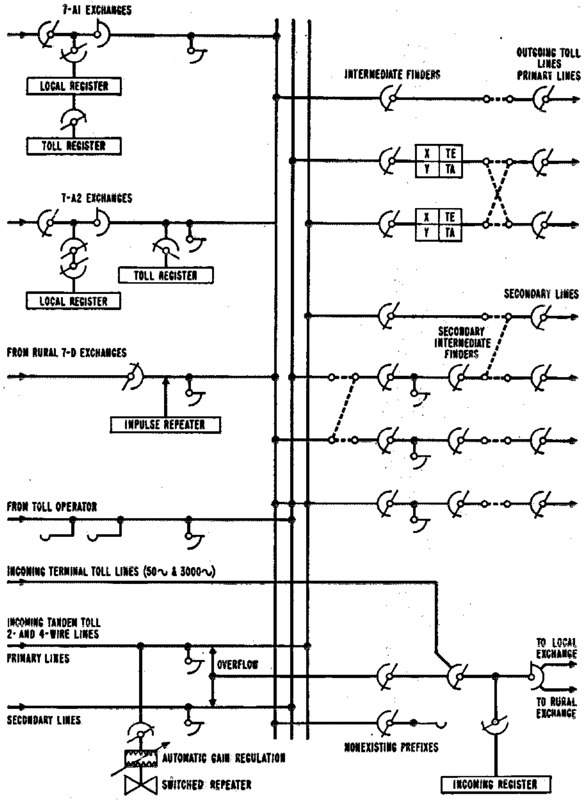

Fig. 4—Simplified junction diagram of the Zurich automatic toll exchange.

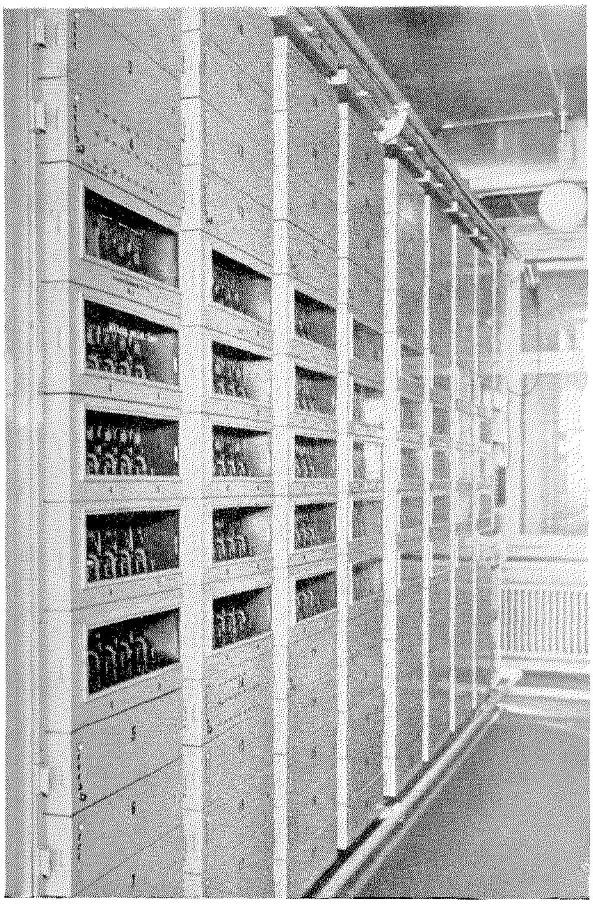

Fig. 5—Selective trunk circuits from 7-A1 local exchanges.

With the dialing of 0 as the first digit, the calling subscriber is extended through a selective trunk circuit to the automatic toll exchange.

3. Selective Trunk Circuits

There are three different classes of selective trunk circuits:

A. Toll traffic originated by city subscribers,
B. Toll traffic originated by rural subscribers,
C. Incoming calls on manual or international toll lines that are to be extended to the automatic toll network.

The selective trunk circuits from the city exchanges are divided into two different classes: incoming from 7-A1, shown in Fig. 5, and from 7-A2 exchanges.

In the 7-A1 exchanges, a toll or auxiliary register is attached to the local register when a subscriber dials 0 as the first digit. The auxiliary register remains attached until all digits are sent out.

For the selective trunks from 7-A2 exchanges, the toll register is located in the automatic toll exchange and the 7-A2 local registers are arranged for sending up to 9 digits into the toll register.

The selective trunks from the local exchanges are combined with a time-and-zone metering circuit and are equipped with three step-by-step selectors of which one serves for the selection of the direction, one for determining the tariff according to the dialed toll prefix, and one for producing a number of artificial tandem selections as may be required.

A tariff-control circuit is also attached to handle networks with mixed tariffs.

The selective trunk circuits for the automatic toll traffic originated by rural subscribers are inserted between the rural main exchange in Zurich and the automatic-toll equipment. During the selection time, an impulse repeater is attached to the selective trunk circuit and is controlled by the auxiliary register in the rural center exchange. This type of selective trunk circuit has no time-and-zone metering circuit because the tariff is determined in the outgoing rural center exchange.

The selective trunk circuit for the toll traffic from the manual toll board is similar to that from the rural network. It appears on the toll-operator positions and also on twin jacks on the cord-circuit repeater positions for connections that require a cord-circuit repeater with automatic gain control. The toll operator dials only two-digit toll prefixes, because the 0 is not required in this case.

There are a number of common circuits used with the selective trunk circuits, such as, master time-control circuit, impulse sender circuit, and delayed back release circuit.

4. Incoming Toll Lines

The incoming toll lines are divided into two classes, which are fundamentally different: terminal lines and tandem lines.

The incoming terminal lines carry only traffic destined for city and rural subscribers of the Zurich group.

The incoming tandem lines are for the transit traffic passing through Zurich to other toll primary or toll end-exchange groups. Overflow facilities are provided so that when all terminal lines are busy, traffic may enter over tandem lines.

The incoming tandem lines are further divided into primary and secondary groups. The primary or main lines carry the incoming traffic from other primary groups, whereas the secondary lines handle traffic from the toll end-exchange groups.

The tandem primary lines are again subdivided into 2-wire and 4-wire lines, which are shown in Figs. 6 and 7.

The 2-wire lines are arranged for 50-cycle signaling, whereas the 4-wire lines, which are used for operation on 12-channel carrier systems, are arranged for 3000-cycle signaling.

The incoming tandem toll lines have access by means of a finder to the switching equipment for repeaters with automatic gain control.

The 2-wire lines are equipped with individual networks.

Each incoming tandem toll line is provided with a switch for connecting the line in the wanted direction, as determined by the digits received from the far end.

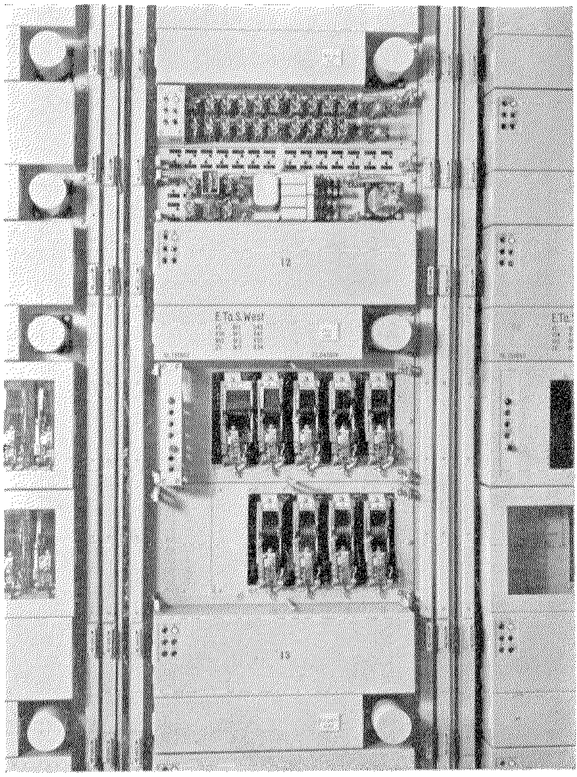

Fig. 6—Incoming 2-wire tandem toll lines.

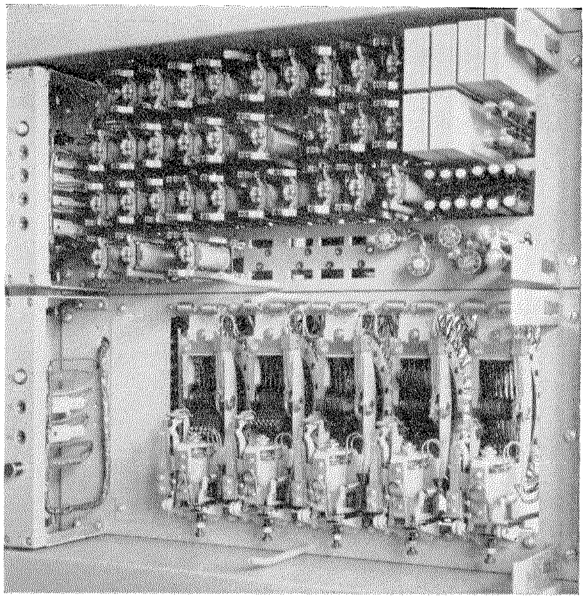

Fig. 7—Incoming 4-wire tandem toll lines.

The incoming tandem lines are connected to the arcs of the intermediate finders via cross-connecting points.

Incoming terminal traffic on tandem lines in case of overflow, is routed through another group of intermediate finders, incoming cord circuits, and incoming registers to the local or rural exchanges, as the case may be.

5. *Intermediate Finder Circuits*

The intermediate finder circuits, shown in Fig. 8, are used for concentrating the traffic between the selecting equipment and the outgoing toll lines. As will be seen from the junction diagram, there are three divisional groups of intermediate finders on the arcs of which are terminated:

A. Selective trunks,
B. Incoming tandem primary lines,
C. Incoming tandem secondary lines.

In the direction of secondary toll centers, there are two stages of intermediate finders. The first intermediate finders include equipment for the selection of the wanted direction. These may be seen in Fig. 9.

The intermediate finders assigned to the incoming tandem toll lines are arranged to provide for overflow from outgoing terminal lines to outgoing tandem lines (X and Y directions) in case all terminal lines in a wanted direction are busy.

A further group of intermediate finder circuits are provided for routing incorrect toll calls for nonexistant toll prefixes to an operator's position.

The intermediate finders are arranged for gradual starting and, to reduce hunting time, starting occurs at the same time the finders of the free outgoing toll lines of the wanted direction are set in motion.

6. *Outgoing Toll Line Circuits*

Each outgoing toll line is terminated on a backward-hunting finder. These circuits, illustrated in Fig. 10, are subdivided into two main groups, viz., primary lines leading to other primary or toll centers and secondary lines connected to secondary or toll end exchanges.

The lines of the first group are again subdivided into terminal and tandem lines for each direction according to the transmission grade. Overflow facilities are provided and, when all terminal lines in a certain direction are busy, the terminating traffic will be routed over tandem lines of the proper direction. This operation is determined by the preceding selective trunking equipment.

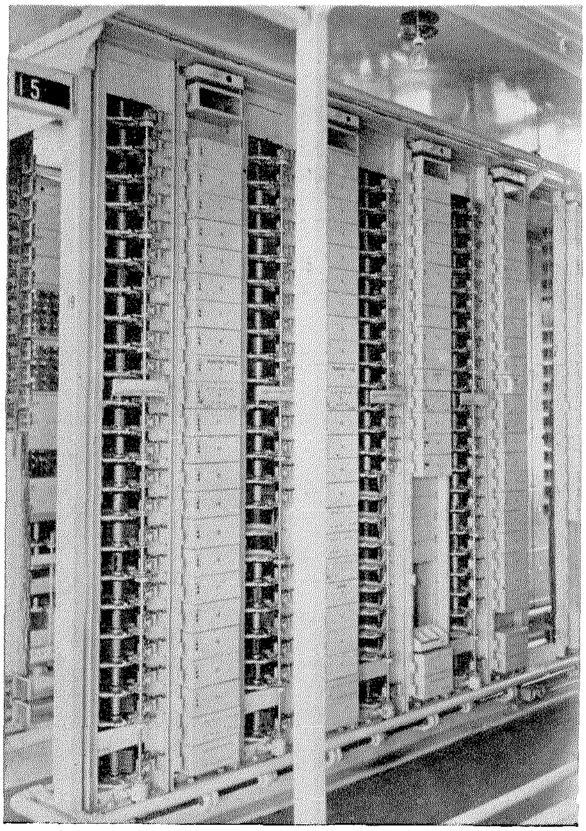

Fig. 8—Intermediate finder equipment.

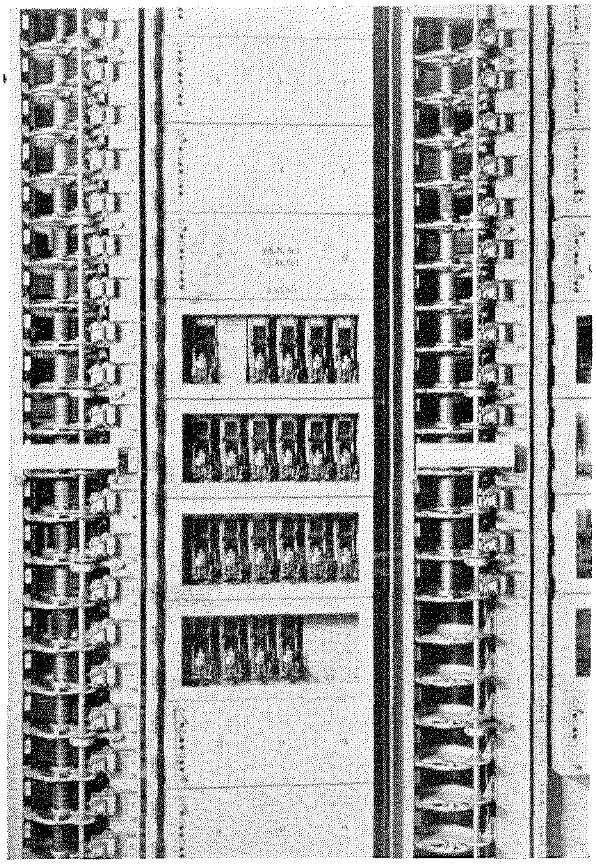

Fig. 9—Intermediate finder circuits with direction markers.

On directions with a great number of lines, gradual starting circuits are provided so that only a limited number of free finders will hunt simultaneously when a certain direction is selected.

The outgoing terminal lines are grouped into *TE1* and *TE2* lines. The *TE1* terminal lines, to which only the outgoing traffic originated by the Zurich city and rural subscribers is routed, are not equipped with networks as such connections never require the insertion of a switched repeater.

The *TE2* terminal lines to which incoming tandem lines also have access are, however, equipped with individual networks in case tandem connections with switched repeaters are to be established. For this purpose, each line circuit is also equipped for line-loss signaling.

The line-loss indication will always be signaled because the circuit does not know where the switched repeaters are connected. This signal is transmitted at the moment the outgoing toll line is connected, i.e., normally at the same time the starting impulse is sent towards the distant exchange.

Direct-current impulses received by the outgoing toll circuit from the local or rural exchanges are translated into alternating-current impulses for transmission to the distant end of the outgoing toll line.

7. *Switched Repeaters and Automatic Gain Control*

Switched repeaters are automatically introduced on tandem connections. Standard 2-wire repeaters have been adapted for this purpose. Each switched repeater circuit is combined with a gain-control circuit consisting of a number of relays and two variable pads for the incoming and outgoing lines. The finder, by which the automatic repeater is connected to the incoming line, consists of 11 brushes of which 2×4 serve for the through connection of the talking and network leads of the incoming and outgoing side.

The repeater connecting and regulating equipment is located in the switch room, whereas the repeaters and gain-control circuits are mounted in the repeater room on the floor below.

The automatic-gain-control circuits may be seen in Fig. 11 and each consists of four variable pads, which can be regulated in 1-decibel steps up to 15 decibels. The switched repeaters are set to give a gain of 17 decibels (1.96 nepers). In the idle position of the repeater, all pads are inserted so that the effective gain is reduced to 2 decibels. As soon as a repeater is switched in, the gain is automatically adjusted to meet the circuit needs.

For setting gain, three distinct indications are required.

A. Loss value of the incoming toll line,

B. Whether the gain of the switched repeater is to be adjusted to compensate for the loss in the incoming toll line or to equal the sum of the loss values of both toll lines, minus 9 decibels (1 neper),

C. Loss in the outgoing toll lines. When the gain-control relays receive a loss indication from the incoming toll line only, the gain of the repeater is set to equal this loss.

On the connections terminating in the Zurich local or rural area, no repeater is used. This is indicated by the setting of the marker switch of the incoming line.

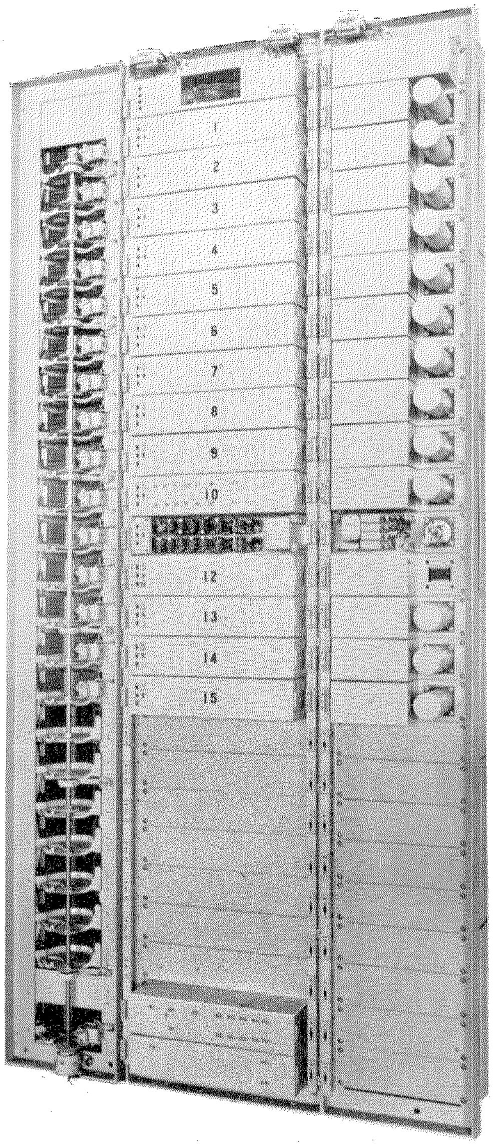

Fig. 10—Outgoing 2-wire toll lines with 50-cycle signaling sets.

8. *Signaling Systems*

Two different systems of signaling are used in the Swiss automatic toll system.

For relatively short 2-wire lines, 50-cycle signaling is used. The signaling current is relayed to bypass the line repeaters.

With the introduction of 12-channel carrier systems in the toll cable network, it became necessary to develop a new voice-frequency signaling system. For this purpose, Standard Telephone and Radio and the Swiss administration jointly have developed a single-frequency 3000-cycle signaling system.

9. *Tandem Connections Over 4-Wire Lines*

With the introduction of 12-channel carrier systems, incoming and outgoing 4-wire line circuits were provided to allow the following through connections:

A. 2-wire to 4-wire lines,
B. 4-wire to 2-wire lines,
C. 4-wire to 4-wire lines.

The incoming and outgoing 4-wire lines are terminated on 4-wire terminating sets.

When 4-wire lines are connected to 2-wire lines, the 3000-cycle signaling impulses incoming on the 4-wire circuit are retransmitted as 50-cycle impulses to the outgoing 2-wire line, and vice versa.

For connections between 4-wire lines, the tail-eating method is used and allows simple switching by avoiding the use of a great number of relay contacts.

10. *Installed Equipment*

The Zurich automatic toll exchange has the following number of principal circuits.

351	Incoming selective trunk circuits from 7-A1 local exchanges,
117	Incoming selective trunk circuits from 7-A2 local exchanges,
100	Incoming selective trunk circuits from 7-D rural exchanges,
20	Incoming selective trunk circuits from manual toll operators,
344	Incoming 2-wire tandem toll lines,
150	Incoming 4-wire tandem toll lines,
460	Incoming 2-wire terminal toll lines,
180	Repeater-connecting and gain-regulating circuits,
1790	Intermediate finder circuits,
432	Intermediate finder circuits with markers,
702	Outgoing 2-wire toll lines,
150	Outgoing 4-wire toll lines.

With the rapid increase of toll traffic in Switzerland, the Zurich automatic toll exchange will be developed further. International toll traffic will undoubtedly introduce new switching problems in automatic operation and these will be the subject of further studies in automatic toll switching.

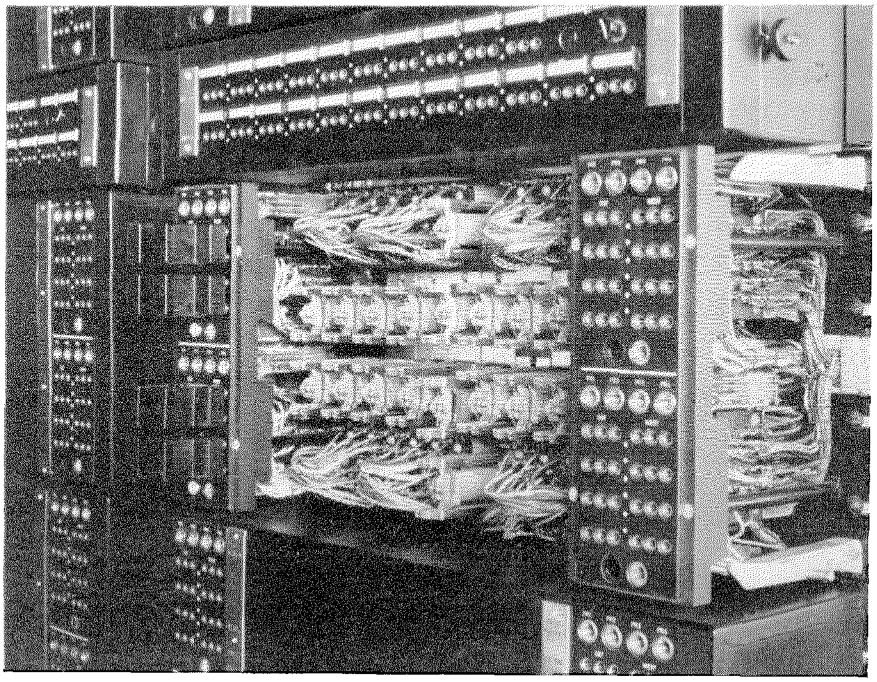

Fig. 11—Pads for automatic gain control of switched repeaters.

Emergency Operation of SOJ-12 Carrier Telephone Systems

By J. M. CLARA and E. SERRA CALABUIG

Compañía Telefónica Nacional de España, Madrid, Spain

PLANT EXTENSIONS for increasing the traffic capacity of the toll routes of Spain by the use of 12-channel open-wire carrier telephone systems were started in 1946. The carrier equipment is the *SOJ*-12 system manufactured by Standard Telephones and Cables, Limited, of London. A description is given of how the circuits on the Madrid–Barcelona route were restored on an emergency basis. This was facilitated by the self-aligning feature of the repeaters and receiving terminals. Although some increase in unintelligible inter-system cross talk would be expected under such conditions, this was not found to be serious enough to prevent commercial use of the circuits.

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During the past ten years, well-known difficulties have delayed telephone development in Spain, where, commencing in 1924, the Compañía Telefónica Nacional de España, then an associate of the International Telephone and Telegraph Corporation, had achieved great progress in the development and quality of its service. Within the limitations imposed by world-wide difficulties, efforts were made to normalize the service and, before the close of World War II, plans had been formulated to provide quick relief for the congested toll service between the most important Spanish cities; this construction was to be started once the war was over.

The toll routes of Spain were totally reconstructed approximately twenty years ago; they have been kept at an extremely high grade of maintenance and, having repaired the damages that resulted from the civil war of our country, at the end of the war we had available a toll plant consisting mostly of open-wire line in first-class condition. It was therefore decided that the first major relief to be given for a period of 3 to 5 years would be based on applying single-, 3-, and 12-channel carrier systems to existing open wire, this having the additional advantage that it required the minimum use of copper, which was in very short supply.

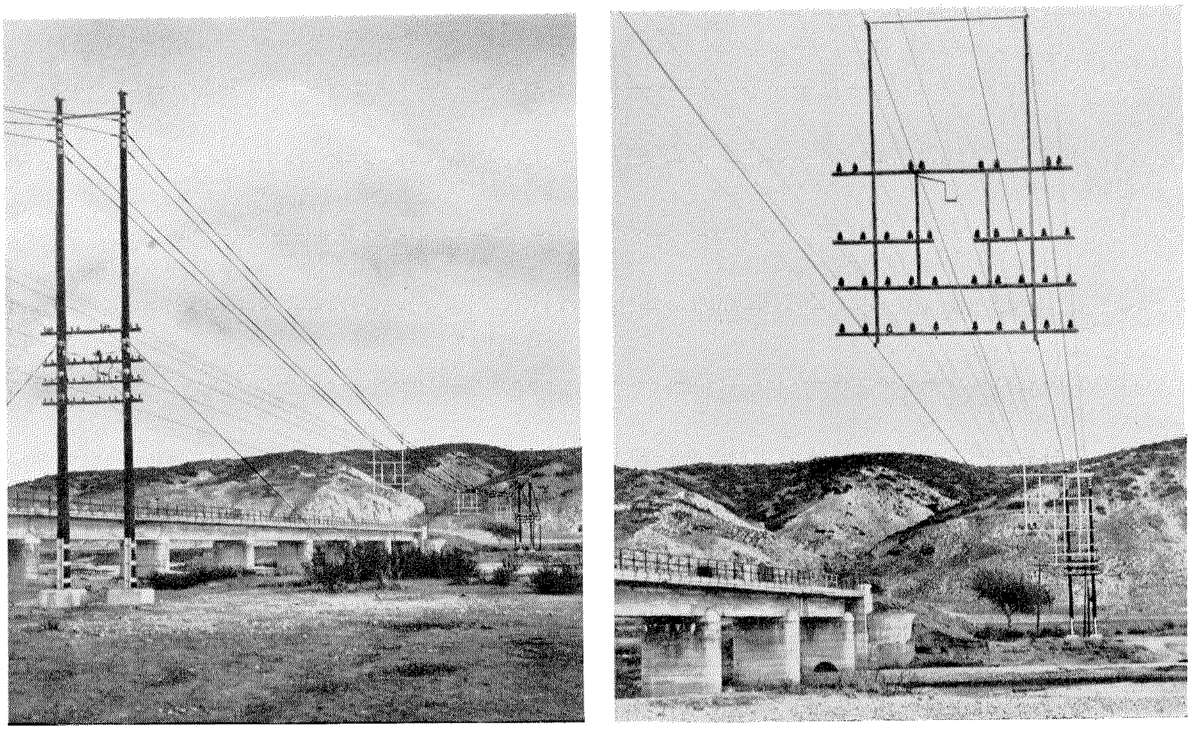
In 1943, a plan was developed for the installation of a number of *SOJ* 12-channel carrier systems, the installation of which is now well advanced. As a description of the over-all plan is beyond the scope of the present paper, we will limit the discussion to experiences during abnormal circumstances where the equipment worked successfully even though the operating margins for which it had been designed were considerably exceeded.

The first route on which *SOJ* 12-channel carrier was applied was Madrid–Barcelona, via Valencia, this having been modified to provide the necessary transpositions so that a number of systems could operate on the same pole line. Three *SOJ* systems were cut into service between October, 1946, and January, 1947.

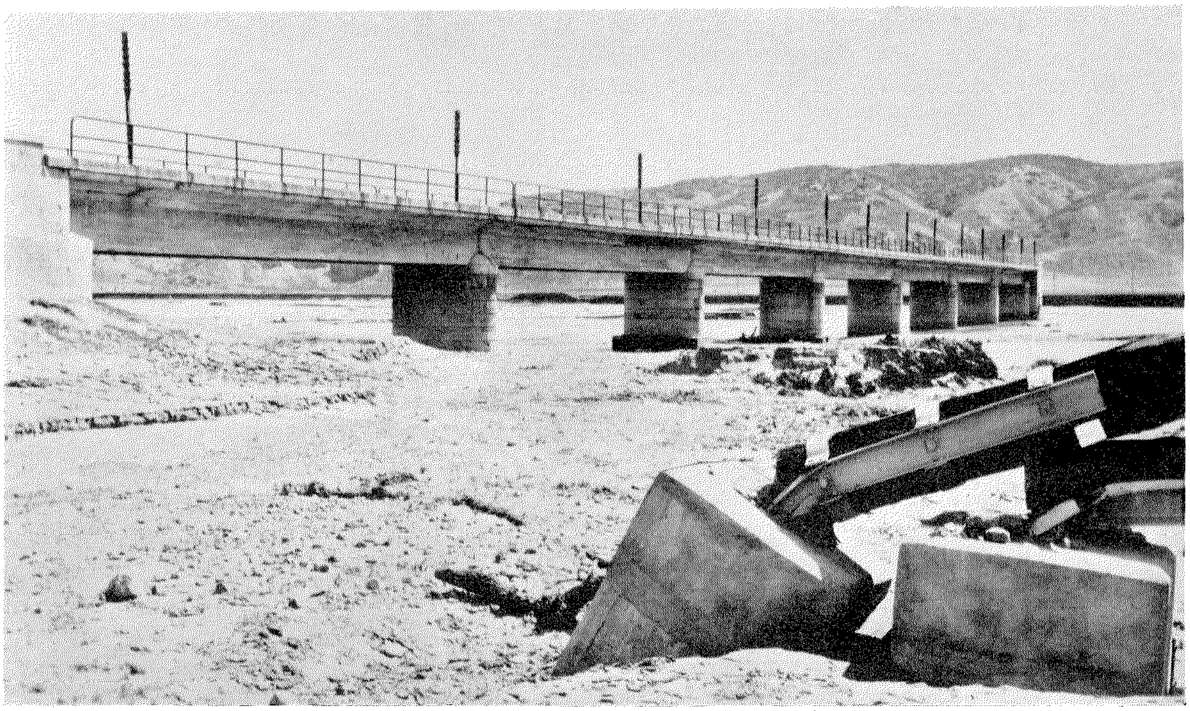
There existed some apprehension that in the event of a major line failure we would be in serious difficulties, since there were no alternate routes for 12-channel carrier operation. Unfortunately, this situation occurred in early March, 1947, when, as a result of exceptionally long-continued rains and melting snows, floods interrupted all types of communication in different parts of Spain.

At the point indicated in the accompanying map, the Madrid–Barcelona route crosses the Jarama River with a 200-meter (656-foot) open-wire span. To obtain correct wire separation and proper transposition spacing, the wires were supported by three crossarm fixtures suspended at 50-meter (164-foot) intervals from messenger strands fastened to special *H* structures at each side of the river. The *H* structures consisted of specially selected tall creosoted pine poles fastened to concrete bases located on the down-river side of the highway approaches to the bridge. In addition to wires, crossarms, and messenger strand, these poles supported a special strand to facilitate line maintenance over the river with the use of a lineman's cable chair.

The rising river flooded the valley and washed out some 200 meters (656 feet) of the eastern highway approach to the bridge, which remained



Left—The Jarama River crossing and the bridge as they originally appeared. The tall *H* structures, shown anchored in concrete bases, were washed down, causing loss of *SOJ* line facilities between Madrid and Barcelona. Right—The frames, which provided for proper spacing and transposition of the line wires, were supported at 50-meter intervals on messenger strands. The bracket in the open space at the middle of the frame, and just below the top crossarm, supported a messenger strand for linemen's cable chairs.



View of the bridge after the flood. Remains of one of the *H* structures are in the right foreground. All land between the end of the bridge and the distant right was washed out. The poles lashed to the bridge railing support the emergency lines.

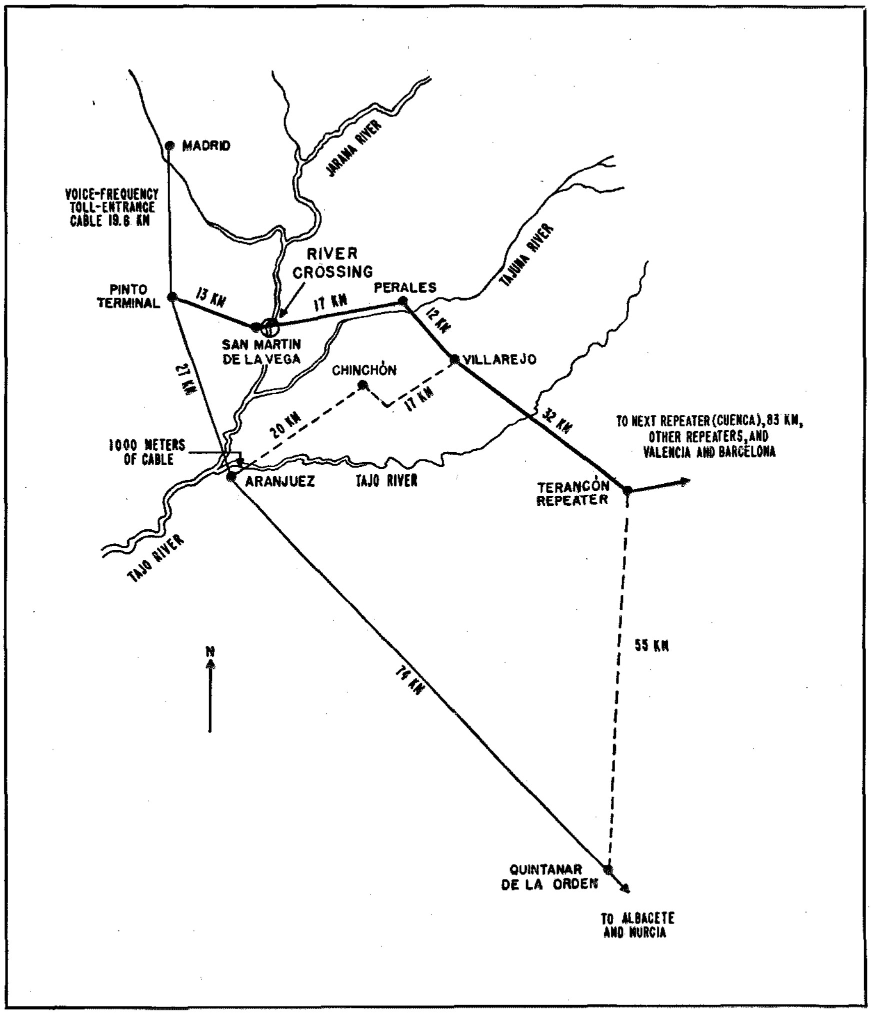
isolated. The force of the flood and the cutting of a new river channel caused the fall of the *H* structure at the eastern side of the bridge, and the line thus fell for a distance of nearly 600 meters (1969 feet).

The first indication of the river-crossing failure was observed at noon on March 5th, when the circuits on the lower arm and two pairs on the top arm were grounded. This put out of operation one of the three *SOJ* systems, and left no spare transposed pairs for carrier. As a trial emergency measure, the out-of-service *SOJ* system was transferred to a Pinto–Cuenca pair of the second arm, and the Tarancon repeater of this *SOJ* system was cut out of service; this latter being necessary because no patching arrangements existed at Tarancon to make use of the circuit for *SOJ*-system operation. In spite of the difference in transmission levels, and the fact that the pair was not transposed for carrier frequencies, the three *SOJ* systems functioned satisfactorily on this basis for seven hours until the rising river washed out the whole crossing.

Consideration was immediately given to the use of some other line facilities for the *SOJ* systems until such time as the normal route could be temporarily repaired. The only two possibilities were the Pinto–Aranjuez–Chinchon–Villarejo route (96 kilometers, or 59.7 miles), and the Pinto–Aranjuez–Quintanar de la Orden–Tarancon route (156 kilometers, or 96.9 miles), in which the Aranjuez–Chinchon–Villarejo and the Quintanar de la Orden–Tarancon sections are secondary or almost rural routes.

The emergency circuits were made up of 1000 meters (3281 feet) of 0.4-millimeter (0.16-inch) subscriber cable, and various sections of 2- and 3-millimeter (0.08- and 0.12-inch) copper open-wire line, partly on brackets.

Patching to obtain through connections was made with rubber-insulated twisted-pair drop wire of 0.8-millimeter (0.32-inch) phosphor-bronze conductor, this being readily at hand. Work was carried out with difficulty as the area was flooded and most highways and roads were cut. Nevertheless, at 7:00 PM the same day,

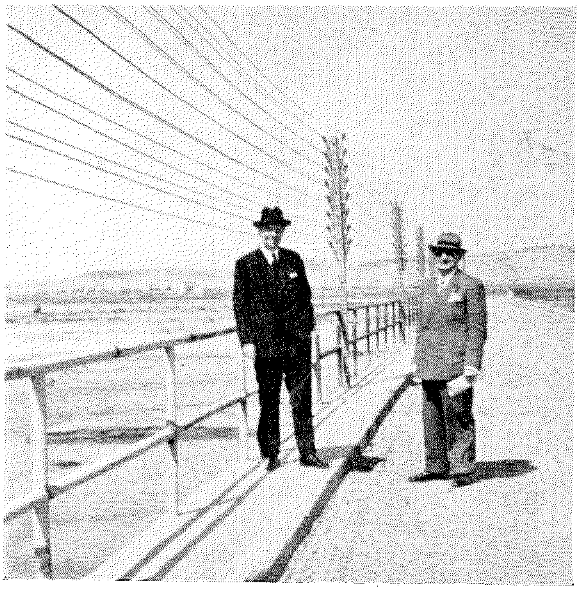


Map of the area in which emergency measures were necessary. The heavy line indicates the routing of the Madrid–Barcelona *SOJ* systems. The light line and dashed lines indicate a secondary or rural route, and the emergency interconnections, respectively.

an *SOJ* and a 3-channel system were placed in operation over the Pinto–Aranjuez–Quintanar de la Orden–Tarancon route, and at 3:00 the following morning, a second *SOJ* and a second 3-channel system were operating over the Pinto–Aranjuez–Villarejo–Tarancon route, thus giving 16 speech channels over each of the two emergency routes. No physical pairs were available for the re-establishment of the third *SOJ* system.

During these rerouting operations, preparations were made for temporary repair of the break at the Jarama River crossing on the normal main route. Light barges were hauled by truck to the nearest bridge approach, and in the early hours of March 6th an attempt was made to cross the river, but this was suspended owing to whirlpools and debris and the swiftness of the current.

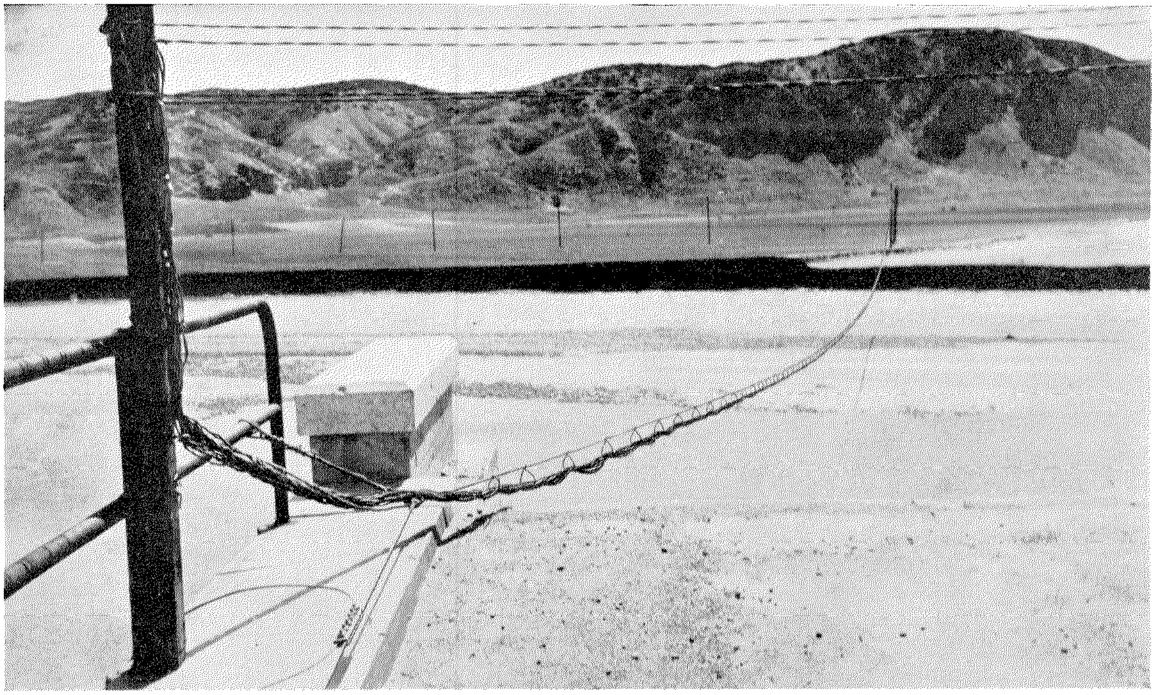
At considerable personal risk and discomfort, line wire, drop wire, and sky rockets were partly carried and partly swum to the Pinto side of the bridge, which had remained standing like an island in the river. From there a line attached to a rocket was shot over the new 200-meter-wide river channel to the Perales side. The rockets used were the largest that could be obtained of the type used in fireworks displays. Once this line was across, drop wires were pulled through this swift current and service re-established. It is interesting to note that with the passing of but one twisted pair across the river, 16 speech circuits (12 *SOJ*-system, one 3-channel-carrier, and one voice-frequency) were obtained. In fact, as the *SOJ* system had been previously turned on and was ready for operation, at the moment of connecting one wire of the river-crossing pair to open-wire temporary termination poles, all channels of the *SOJ* system commenced to operate even though the other wire of the circuit had not yet been spliced, so demonstrating one of the peculiarities of this high-frequency carrier equipment.



View of the emergency lines on crossarms with diagonal brackets. The authors are in the foreground.

The emergency wire used in this temporary repair consisted of rubber-covered twisted pair with 0.8-millimeter phosphor-bronze conductor. Approximately 500 meters (1640 feet) of this wire were spliced into each pair of the damaged section, half the length being fastened to the side of the bridge and the other half supported in a ring run suspended from a messenger cable strung over the new river channel; the river span was too wide for the drop wires to be self supporting. For the carrier pairs, one twisted pair was used for each conductor, these being separated from each other and the other wires by about 15 centimeters (6 inches) in the 250-meter (820-foot) section along the bridge. Separation between wires and circuits was, of course, impossible in the ring run over the new river channel. As each of these pairs were re-established, the three *SOJ* systems, as well as the other carrier systems, were transferred from their temporary rerouting back to this route; all functioned satisfactorily in spite of the impedance irregularity and close physical association of the wires at the crossing.

The permanent rebuilding of the Jarama River crossing will take considerable time, and once the river has completely subsided, a provisional reconstruction is proposed. However, at the time of this writing (May, 1947), after two months of operation over the emergency river crossing, the three *SOJ* and four 3-channel carrier systems continue functioning without any abnormalities, and with scarcely any impairment in quality of service. It will, of course, be necessary to reinstate the affected line section to its former standards of construction in order to ensure continued and satisfactory service and to maintain the previous standards of low inter-



Ring run over the new 200-meter channel. In spite of the loss of wire spacing in each circuit, and the close proximity of all circuits, the *SOJ* and 3-channel systems operated satisfactorily.

system cross talk under varying conditions of line attenuation with changes in weather.

Our former fears as to difficulties which might be encountered in the use of *SOJ* 12-channel systems in the event of a major line break not only have been dissipated, but we have also found that the characteristics of the 12-channel carrier equipment make it possible to re-establish a large number of communication channels over damaged line facilities with a speed difficult if not impossible to obtain where only voice-frequency equipment is available.

Nickel–Iron Alloy Dust Cores*

By S. E. BUCKLEY

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THE DEVELOPMENT of dust cores to meet the increasing demand for use in telecommunications equipment is outlined. Requirements for such uses are defined, a high and constant permeability and low losses in the core being the most important. Amongst the materials used for dust cores described are electrolytic iron, Permalloy (nickel-iron), and molybdenum–Permalloy.

The constancy of permeability and its influence on design, the question of calculating the dust-core permeability from inherent properties of the metal, and core losses are discussed; some figures are given illustrating the result of reducing core-loss coefficients over the course of dust-core development.

The effective resistances at audio frequencies of dust-cored coils of various materials are compared.

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1. *Performance Requirements of Magnetic Dust Cores*

The first large-scale uses of ferromagnetic materials in electrical engineering were principally in equipment for the generation and distribution of electrical power. In these applications, the use of magnetic material in sheet form in the construction of generators and transformers has been common practice for many years. With the continuous development of telephony and telegraphy, and related telecommunications systems, the scope of use of ferromagnetic materials was greatly increased. The performance required of magnetic materials in telecommunications equipment covers a very wide range and is often of a very specialized nature and the problems encountered are considerably more complex than those in power engineering. In this paper, the manner in which dust cores, more especially nickel–iron alloy dust cores, have been developed to meet some of these special requirements will be considered.

The general requirements for a magnetic dust core are as follows. The core permeability must be reasonably high and must remain constant over a wide range of induced flux density, and also over a wide range of frequency. The permeability must remain constant with time, and the variations due to temperature changes and to accidental overloads must be small. The total losses occurring in the core must be low, and in many instances it is important for the hysteresis loss to be especially small. The process of manufacture has to be controlled so that the various requirements are all met within narrow limits of tolerance. In addition to meeting the electrical performance required, the cores must be sufficiently robust to withstand the subsequent handling entailed in such operations as coil winding, adjusting, impregnating, and potting.

The manner in which the core performance requirements led to the development and commercial production of dust cores may be stated briefly. In order to reduce eddy-current losses in the core, subdivision of the magnetic material in a direction perpendicular to that of the magnetic flux is necessary. The practical achievement of this requirement is illustrated by core constructions in which the magnetic material takes the form of thin insulated laminations or fine wire. Although this construction results in a reduction of eddy-current losses it does not ensure constancy of permeability under the various conditions already mentioned. To obtain this constancy, the magnetic material must be subdivided to introduce air-gaps in a direction perpendicular to the plane of lamination; for example, in a ring core, radial air-gaps are necessary. Thus, for some applications, the core would need to be assembled from spaced sectors of material, the sectors consisting either of stacks of insulated thin laminations, or else of strands of insulated thin wire bonded together. This general type of construction is cumbersome and costly, although it was used at one time for

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* Reprinted from "Symposium on Powder Metallurgy," The Iron and Steel Institute, London, England, pp. 59–63; May, 1947.

loading-coil cores for telephone lines. A more practical commercial solution to the problem was sought.

2. *Outline of Progress in Core Development*

It had been appreciated for a long time, as is shown by a suggestion made by Heaviside,[1] in 1887, that a core composed of finely divided iron insulated, for example, with wax might exhibit the desired properties. Although cores of this type can have low losses and other desirable characteristics, the permeability obtainable is very low. It was not until the advantages of using forming pressures of the order of 100 tons per square inch were demonstrated, that the large-scale production of high-permeability dust cores was achieved. This development by the Bell Telephone Laboratories[2] in America was given impetus during the first World War, when the supply from Europe of diamond dies for the drawing of fine iron wire was stopped, and hence the production of iron-wire cores was jeopardized.

3. *Electrolytic-Iron Dust Cores*

The outline of the core technique then developed was as follows. Pure iron was deposited electrolytically under conditions which yielded a brittle product. The brittle iron was ball-milled to produce iron powder, which was next coated successively with zinc and shellac, and then pressed at about 100 tons per square inch into cores of permeability 35. The cores thus obtained had losses which were low enough for the cores to be used in loading (Pupin) coils for insertion in telephone-cable circuits and in other high-quality coils included in telephone-circuit equipment. The type of core which had previously been used consisted of fine iron wire wound into toroidal form and then cut radially to introduce stabilizing air gaps. The performance of the dust cores was greatly superior to that of the wire cores.

4. *Permalloy Dust Cores*

After the publication, in 1923, of Arnold and Elmen's[3] paper on Permalloy, the magnetic properties of nickel–iron alloys attracted considerable attention. It had been demonstrated that after suitable treatment, including especial care in heat treatment, these alloys exhibited much higher permeabilities and yet a much lower hysteresis loss than previously known materials. Since the nickel–iron alloys in sheet form had such excellent properties, the possibility of obtaining improved dust cores by using the alloys in dust form was apparent. Various methods of producing alloy powders have been suggested. However, it is important to remember that the magnetic properties of a metal or alloy, in common with many other physical properties, are determined by the crystalline structure of the material as well as by its chemical composition. Thus, in many instances, a material which exhibits good magnetic properties when in the solid form may be reduced to powder but, owing to the treatment involved in the powdering process, the magnetic properties are completely changed. Hence the process must be developed to produce powder which retains as far as possible the inherent magnetic properties of the solid material. A number of processes were investigated for the production in powder form of nickel–iron alloys containing approximately 80 per cent of nickel. The first of these methods for large-scale production is described in a paper by W. J. Shackelton and I. G. Barber,[4] published in 1928. Nickel and iron are melted together and a controlled amount of sulphur is allowed to remain in the melt so that when the ingots are subsequently hot-rolled, and then allowed to cool, a thin layer of sulphide separates out at the grain boundaries. The relative proportions of manganese and sulphur in the material must be so arranged that the alloy is sufficiently malleable to roll while hot, and yet sufficiently brittle on cooling for slabs 1/4 inch in thickness to be readily broken up. Fig. 1 is a photomicrograph showing the polygonal crystals separated by the sulphide layer. On pulverizing the slab, disintegration occurs at the grain boundaries so that the particles of powder correspond generally to the crystals seen in the slab. Fig. 2 shows the appearance of the particles and confirms how they resemble the grains in the structure shown

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[1] O. Heaviside, *The Electrician*, p. 302; February 11, 1887.

[2] B. Speed and G. W. Elmen, *Transactions of the American Institute of Electrical Engineers*, v. 40, p. 596; 1921.

[3] H. D. Arnold and G. W. Elmen, *Journal of the Franklin Institute*, v. 195, n. 5, p. 621; 1923.

[4] W. J. Shackelton and I. G. Barber, *Transactions of the American Institute of Electrical Engineers*, v. 47, p. 429; 1928.

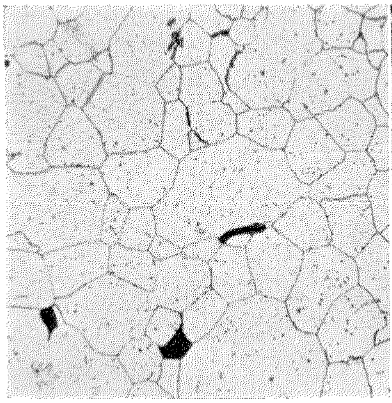

Fig. 1—Section of brittle molybdenum–Permalloy.

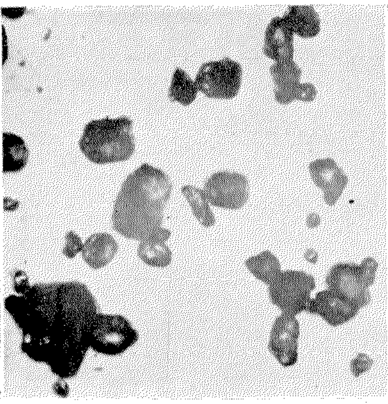

Fig. 2—Molybdenum–Permalloy dust.

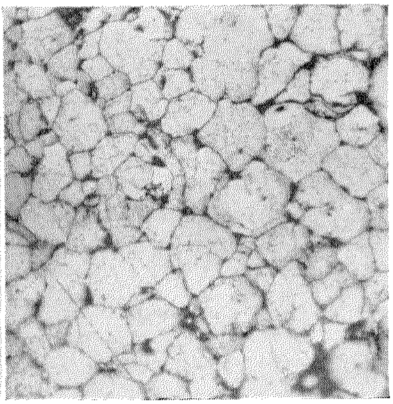

Fig. 3—Section of pressed molybdenum–Permalloy dust core.

in Fig. 1. Since the alloy has been work-hardened in the pulverizing process, it is annealed before proceeding to the stages of insulating and pressing. During the anneal some sintering occurs, but the original state of fineness is restored by light repulverizing. The annealed powder is next insulated by coating the particles with a thin layer of ceramic-type material which must withstand the subsequent pressing and core-annealing operations without rupturing. The insulated powder is pressed into cores at pressures up to 100 tons per square inch. The cohesion of these cores is derived mainly from the interlocking of the particles but, as already stated, the insulating layer must withstand this distortion during pressing. The high pressure which is necessary to produce close packing of the particles adversely affects the magnetic properties of the core, so that annealing of the pressed core is necessary in order to develop the desired properties. From powder containing approximately 80 per cent of nickel, dust cores of permeability 100 were produced. Although this permeability was much higher than that of the electrolytic-iron dust cores, the losses in the alloy dust cores were considerably lower, thus enabling a considerable reduction in core size to be made for a given coil performance.

5. Molybdenum–Permalloy Dust Cores

Further development of alloys of the Permalloy type in sheet form led to the development of nickel–iron alloys containing small percentages of other elements.[5] Thus 3.5 molybdenum–Permalloy, containing 3.5 per cent of molybdenum, has a higher initial permeability than the straight nickel–iron alloy, and has the additional advantages of lower hysteresis and lower eddy-current losses. Moreover, the heat treatment of molybdenum–Permalloy is much simpler and less critical than that of the straight alloy. The same general principles that were used for making 80-per-cent nickel–iron alloy dust cores were applied to making cores of 2:81 molybdenum–nickel–iron.[6] A number of improvements in the process were introduced and cores of permeability 125 were developed. Despite the higher permeability, the characteristics of these cores are superior in all respects to those of the previous type of permeability 100.

A section of a pressed dust core is shown in Fig. 3. Comparison with Figs. 1 and 2 shows the progressive stages by which the grains of the original solid material proceed through pulverizing to the final insulated and pressed state. The average particle size in the core is 0.002 inch.

6. Constancy of Permeability of Dust Cores

Fig. 4 compares the change of permeability with applied magnetizing force of sheet materials with corresponding dust cores. There has been considerable discussion as to the calculation of dust-core permeability from the inherent properties of the material.[7] Although the matter is not yet finally settled, the following elementary treatment gives a good illustration of the effect, and claims to be no more than an illustration.

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[5] G. W. Elmen, *Journal of the Franklin Institute*, v. 207, p. 583; 1929.

[6] V. E. Legg and F. J. Given, *Transactions of the American Institute of Electrical Engineers*, v. 59, p. 865; 1940.

Let the relative length of magnetic particles to insulating material effective in the direction of the magnetic flux in the core be $l:a$. Let the permeability of the solid material be μ_1 and the permeability of the dust core be μ. Then, by the usual reluctance relationship for the magnetic circuit,

$$\frac{l+a}{\mu}=\frac{l}{\mu_1}+a.$$

From this,

$$\mu=\frac{\mu_1(l+a)}{l+a\mu_1}.$$

If the effective ratio of $l:a$ is 100:1, the following corresponding values of solid-material and dust-core permeabilities would result:

Solid Material:	1000	10,000	100,000
Dust Core:	92	100	101

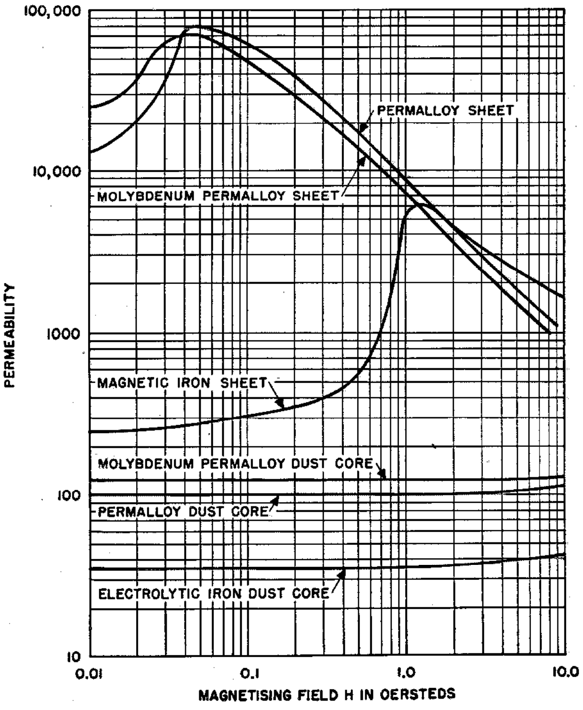

Fig. 4—Variation of permeability with magnetizing field for sheet and dust-core materials.

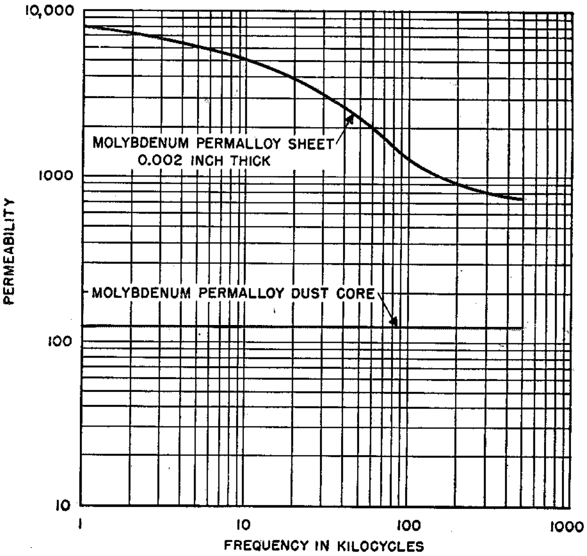

Fig. 5—Variation of permeability with frequency of thin molybdenum–Permalloy tape and dust core.

As already stated, the treatment given here is approximate only, and for further consideration the reference already quoted may be consulted, but the general manner in which the gap ratio has a stabilizing effect on the permeability of the dust core is demonstrated.

Similarly, Fig. 5, which compares the change with frequency of the permeability of thin tape material with that of a dust core, illustrates the stabilizing effect of the gaps.

7. *Core Losses of Dust Cores*

In electrical power engineering it is usual to express the magnetic core losses in the form of watts loss per pound or per kilogramme. In communications engineering, interest is usually centred on the effective resistance in ohms of a coil, so that an expression of core losses in the form ohms per henry (inductance) is generally used. It has been established[8, 9] that the core loss of a dust core, or of a sheet magnetic material at low flux densities, may be expressed in the form

$$\frac{\Delta R}{\mu L}=(aB_m+c)f+ef^2,$$

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[7] G. W. O. Howe, *Wireless Engineer*, v. 23, p. 291; November, 1946.

[8] V. E. Legg, *Bell System Technical Journal*, v. 15, p. 39; 1936.

[9] H. Jordan, *Elektrische Nachrichten-Technik*, v. 1, p. 7; 1924.

where ΔR = core loss in ohms,
μ = core permeability,
L = inductance in henries,
f = frequency in cycles per second,
a = hysteresis coefficient of core loss,
c = residual coefficient of core loss,
e = eddy-current coefficient of core loss,
B_m = maximum flux density induced in the core.

The progress which has been made in reducing core-loss coefficients in the successive stages of dust-core development is shown in Table I.

TABLE I

Material	Permeability	Hysteresis Loss, $a\times10^6$	Residual Loss, $c\times10^6$	Eddy-Current Loss, $e\times10^9$
Electrolytic-iron dust	35	49	109	88
80:20 Permalloy dust	100	5.0	60	30
2:81 molybdenum–Permalloy dust	125	1.2	32	22

The reduction of hysteresis and residual losses is due to improvements in the magnetic quality of the materials. The reduction of eddy-current loss arises from higher electrical resistivity of the alloys.

8. *Resistance Characteristics of Dust-Cored Coils*

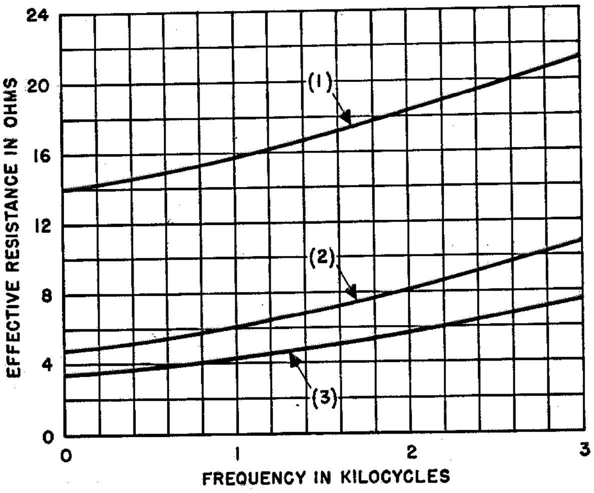

Fig. 6—Comparison of effective resistance at audio frequencies of dust-cored coils. All cores are 30 cubic centimetres in volume; coil inductance = 100 millihenries. (1), Electrolytic iron dust, permeability = 35; (2), 80:20 Permalloy dust, permeability = 100; (3), 2:81:17 molybdenum-Permalloy, permeability = 125.

Fig. 6 shows the relative values of effective resistance at audio frequencies of coils wound on equal-sized cores (30 cubic centimetres) of the different materials. The great improvement obtained by using Permalloy cores is very apparent. It would be possible to design a coil on an electrolytic-iron dust core to have a direct-current resistance equal to that of the molybdenum–Permalloy dust-cored coil, but this would require the iron-dust core to be almost ten times as large. However, apart from direct-current resistance there are other considerations of very great importance. For many coil designs it is essential to restrict the hysteresis loss to a very low value. For equal hysteresis loss, the volumes of an iron-dust core and a molybdenum–Permalloy dust core would be in the ratio of 30:1. In practice, this means that in some applications it is now possible to use a dust-cored coil where previously no reasonable design was available. The usual procedure in replacing coil designs in the improved materials has been to produce improved coil performance combined with reduced size.

9. *Influence of Core Permeability on Coil Design*

Fig. 7 illustrates the resistance characteristics of coils wound on 30-cubic-centimetre cores of molybdenum–Permalloy of two permeabilities, 125 and 50. It will be noted that although the core losses of the 50-permeability core are ap-

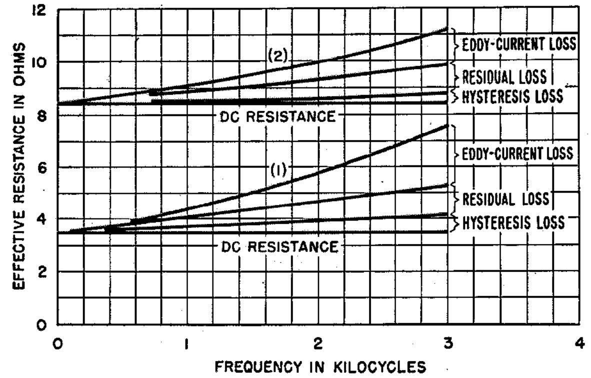

Fig. 7—Effective resistance at audio frequencies of molybdenum–Permalloy dust-cored coils. All cores 30 cubic centimetres in volume; coil inductance = 100 millihenries. (1), core permeability = 125; (2), core permeability = 50.

preciably lower than those of the 125-permeability core, the effective resistance of the former coil is considerably higher over the frequency range shown. The higher value is mainly due to the higher direct-current resistance. In order to reduce this resistance to that of the other coil, the core size would have to be increased by four times. This example shows the great importance of high-permeability cores of low loss in obtaining low-resistance coils of small size for audio-frequency use.

It is apparent from Fig. 7 that, although the two curves do not cross below 3 kilocycles, they will cross at a higher frequency above which the lower-permeability coil will have the better characteristics. Thus, when it is required to produce coils for use at various frequencies, it becomes necessary to produce cores of more than one permeability in order to cover the frequency range adequately. A measure of coil quality, that is often used and that enables the comparison of core materials to be shown conveniently on one graph for a wide frequency range, is the Q factor. The Q of a coil is the quotient of the coil reactance divided by the effective resistance, thus, $Q=\omega L/R$, and a high value of Q denotes a good coil. Fig. 8 shows the variation of Q with frequency for coils wound on 30-cubic-centimetre cores of various permeabilities. The magnetic material used throughout was molybdenum–Permalloy, and the general principles of coil manufacture are the same, the various permeabilities being obtained by different degrees of dilution of the core with nonmagnetic material. It will be seen that the 125-permeability core is best in the range up to 5 kilocycles, the 50-permeability core for the range 5–15 kilocycles, the 25-permeability core for 15–50 kilocycles, and the 13-permeability core from 50 kilocycles upwards.

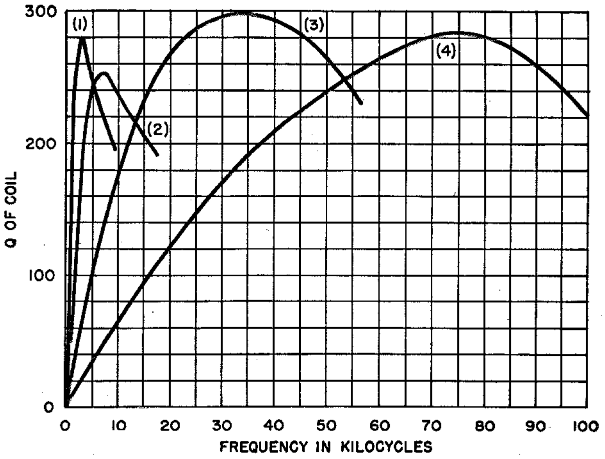

Fig. 8—Q values of coils on molybdenum–Permalloy dust cores of various permeabilities. All cores 30 cubic centimetres in volume; coil inductance range = 10–20 millihenries. Core permeabilities: (1) = 125, (2) = 50, (3) = 25, and (4) = 13.

10. Summary

This paper deals essentially with the properties of nickel–iron alloy dust cores made by the methods described. It is appreciated that there are other methods[10] by which alloy dust cores may be produced, and some information on these methods is already published. There is also a considerable range of low-permeability magnetic-dust cores produced by methods based on the general principles of plastic moulding. Information on these cores, which are widely used at radio frequencies, is also available from publications.

11. Acknowledgment

Thanks are due to Mr. E. L. Bodycombe for the photographs shown in Figs. 1, 2, and 3.

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[10] G. R. Polgreen, *Post Office Electrical Engineers' Journal*, v. 37, p. 1; April, 1944.

Very-High-Frequency Single-Channel Receiver

By W. C. LANE and T. C. CLARK

Federal Telephone and Radio Corporation, Newark, New Jersey

INCREASED traffic at the major airports and the greater coverage required of airline enroute communications have shown the need for an improved very-high-frequency crystal-controlled ground-station receiver with high rejection of unwanted signals and good sensitivity. The receiver designed for these requirements operates on a single pretuned frequency in the range of 118 to 136 megacycles per second, and employs a 14-tube superheterodyne circuit, incorporating automatic-volume-control, squelch, and noise-limiter circuits. Low cost, small size, high quality, performance, and an easily serviceable unit were the points stressed in this design. Field tests indicate the effectiveness of the design.

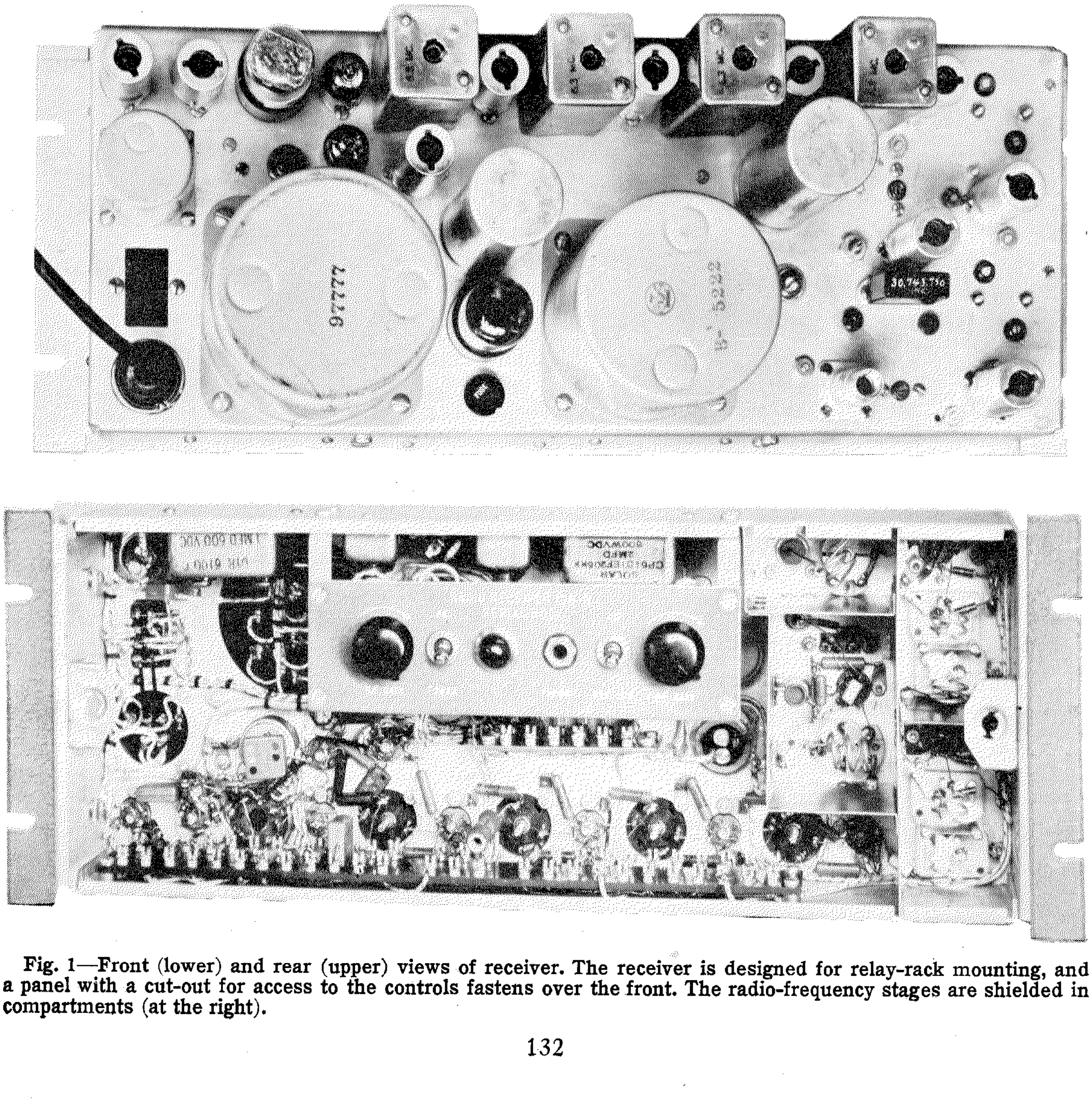

Fig. 1—Front (lower) and rear (upper) views of receiver. The receiver is designed for relay-rack mounting, and a panel with a cut-out for access to the controls fastens over the front. The radio-frequency stages are shielded in compartments (at the right).

1. Physical Characteristics

Mechanically, the receiver will mount in any standard relay rack and, since tubes and transformers are mounted horizontally at the rear, tube replacements and all tuning adjustments can be made with a minimum of effort. The front panel is removable (Fig. 1) to facilitate inspection and repair without taking the receiver from the rack. Operating controls are mounted on a subpanel immediately behind a cutout in the front panel, and include an audio-frequency gain control, radio-frequency gain control, power on-off switch, squelch on-off switch, pilot lamp, and phone jack. Fig. 1 shows the unit-type construction employed.

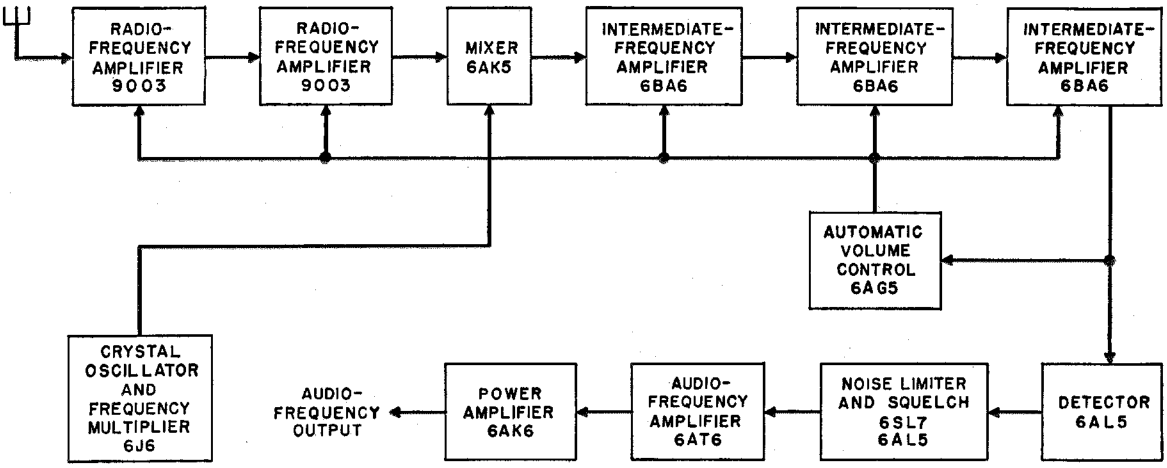
RADIO-FREQUENCY AMPLIFIER 9003
RADIO-FREQUENCY AMPLIFIER 9003
MIXER 6AK5
INTERMEDIATE-FREQUENCY AMPLIFIER 6BA6
INTERMEDIATE-FREQUENCY AMPLIFIER 6BA6
INTERMEDIATE-FREQUENCY AMPLIFIER 6BA6
AUTOMATIC VOLUME CONTROL 6AG5
CRYSTAL OSCILLATOR AND FREQUENCY MULTIPLIER 6J6
AUDIO-FREQUENCY OUTPUT
POWER AMPLIFIER 6AK6
AUDIO-FREQUENCY AMPLIFIER 6AT6
NOISE LIMITER AND SQUELCH 6SL7 6AL5
DETECTOR 6AL5

Fig. 2—Block diagram of single-channel 118–136-megacycle receiver. Squelch and noise-limiter stages improve the usefulness at airports, where there is generally considerable electrical noise.

2. Electrical Characteristics

Every effort was made in the design to use conventional and straightforward circuits in achieving the high-performance characteristics required. Fig. 2 is a block diagram of the receiver. The sensitivity of the prototype model was 0.7 microvolt for 50-milliwatts output at a 10-decibel signal-to-noise ratio. The image response was found to be down 86 decibels at the high-frequency end of the band, and the spurious responses resulting from oscillator harmonics were down 94 decibels or more.

2.1 RADIO-FREQUENCY STAGES

Because of the relatively low intermediate frequency, which was necessary to achieve the required selectivity, it was found desirable to incorporate two double-tuned radio-frequency stages and a tuned antenna stage to attain high image rejection. The coils are constructed of 14-American-Wire-Gauge silver-plated copper wire, and are self-supported on the variable air capacitors for the shortest lead lengths and to obtain the highest Q. The 9003-type tube is chosen for radio-frequency amplifier, partly because of its remote-cut-off characteristic, which makes it readily adaptable to automatic-volume-control application, and also because of the small input loading. As an indication of the importance of input loading at these frequencies, two 9003's and a 6AK5 (used as amplifiers and mixer, respectively) were used to replace two 6BA6's and a 6AG5. With the tubes biased to the same transconductance and with the same coupling coefficient, the gain then increased 6 decibels and the image rejection improved 15 decibels.

Small baffle shields isolating the plate and grid pins at the tube sockets and a high degree of shielding between stages are necessary to minimize regeneration, and this allows an over-all radio-frequency gain of 140 to be attained.

The bias on the radio-frequency tube grids is kept at the lowest permissible value to obtain the optimum signal-to-noise ratio.

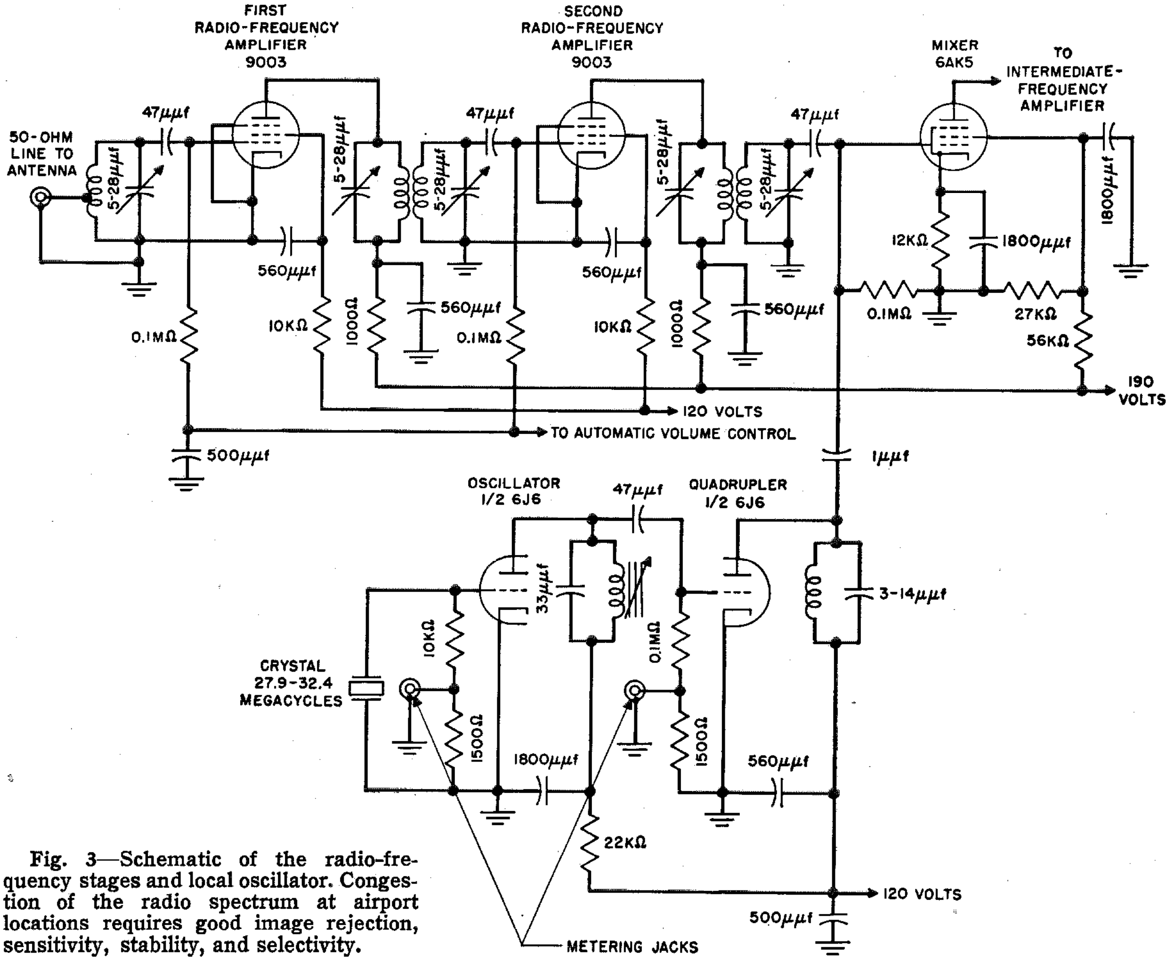

Fig. 3—Schematic of the radio-frequency stages and local oscillator. Congestion of the radio spectrum at airport locations requires good image rejection, sensitivity, stability, and selectivity.

2.2 Oscillator, Multiplier, and Mixer

The oscillator circuit (Fig. 3) is of the tuned-plate tuned-grid type; insertion of the proper crystal in the grid circuit will permit oscillation between 27.9 and 32.4 megacycles. The third-harmonic-type crystal is mounted in an unheated *CR-1* holder and has an over-all frequency tolerance of ±0.005 percent between temperature limits of -25 and $+75$ degrees centigrade (-13 and $+167$, Fahrenheit).

A third-mode crystal, with a built-in thermal unit to keep the crystal at a constant temperature, was employed on a previous model. This did not prove successful, since the opening and closing of the heater contacts set up radio-frequency noise currents that were amplified by the sensitive receiver to produce objectionable noise output.

A twin-triode tube, type 6J6, is used for the oscillator and multiplier. The crystal frequency is multiplied four times in a single stage, and is applied to the mixer grid through a 1-micro-microfarad coupling capacitor. Because of the high fundamental crystal frequency involved, the undesired third and fifth harmonics generated in the oscillator and quadrupler, and which are present at the mixer grid by an amount dependent on the quadrupler plate-circuit selectivity and the crystal activity, are far removed in frequency from the desired fourth harmonic. Thus, with the aid of the front-end selectivity, the spurious responses due to oscillator harmonics were practically eliminated. The oscillator fundamental and harmonic radiation is kept to a low value by proper shielding of the oscillator and quadrupler stages, and by filtering the power-supply leads to the compartment with button-mica feed-through capacitors. The radiation from the prototype model, as measured across the 50-ohm input circuit, was strongest at a frequency of 129.675 megacycles, for the channel frequency of 136 megacycles, and this was only 100 micromicrowatts.

The 6AK5 mixer operates with only 50 volts on the screen grid so that the injection voltage from the quadrupler may be small, but sufficient conversion gain may be obtained. The percentage of noise introduced by the mixer tube is negligible due to the high amplification preceding it.

2.3 INTERMEDIATE-FREQUENCY AMPLIFIER

The intermediate-frequency amplifier design was dictated by several considerations of selectivity: Present frequency allocations assign channels 200 kilocycles apart, so with the allowable combined drift of the transmitter and receiver, the bandwidth at the 6-decibel attenuation points must not be less than 48 kilocycles. In the event of 100-kilocycle channeling, the selectivity curve should afford at least 60 decibels of discrimination against an adjacent channel, which requires that the bandwidth at 60 decibels attenuation be no more than 180 kilocycles. By incorporating four double-tuned transformers in three 6.325-megacycle stages, the selectivity curve shown in Fig. 4 is obtained. A shape factor (which is the ratio of the bandwidth at 60 decibels to the bandwidth at 6 decibels) of 2.88 is indicated. Three remote-cut-off 6BA6 tubes are used in this system.

With the limitations imposed by present and possible future channel assignments, it is important that the intermediate-frequency system have a greater frequency stability than the preselector with regard to shifts of center frequency and bandwidth with changes in temperature, bias, line voltage, etc. In the design of this amplifier, iron-dust-core transformers with fixed silver-mica capacitors have been incorporated, and a high degree of temperature stability has been obtained. The center-frequency shift of the 6.325-megacycle intermediate-frequency system with a change in temperature from 0 to 40 degrees centigrade (+32 to +104 degrees Fahrenheit) is found to be less than 5 kilocycles. Due to the large tuned-circuit capacitance of 250 micromicrofarads, the Miller effect with gain-control bias variations is minimized and the center frequency shifts less than 4 kilocycles with a 60-decibel change in gain and a variation of ±10 percent in line voltage.

2.4 AUTOMATIC-VOLUME-CONTROL SYSTEM

The output of the automatic-volume-control amplifier is designed to regulate the grid bias of the two radio- and three intermediate-frequency stages in order that the output of the receiver will remain essentially constant with inputs between 2 microvolts and 0.1 volt (Fig. 5). Below the threshold value, the automatic-volume-control amplifier is cut off, so that

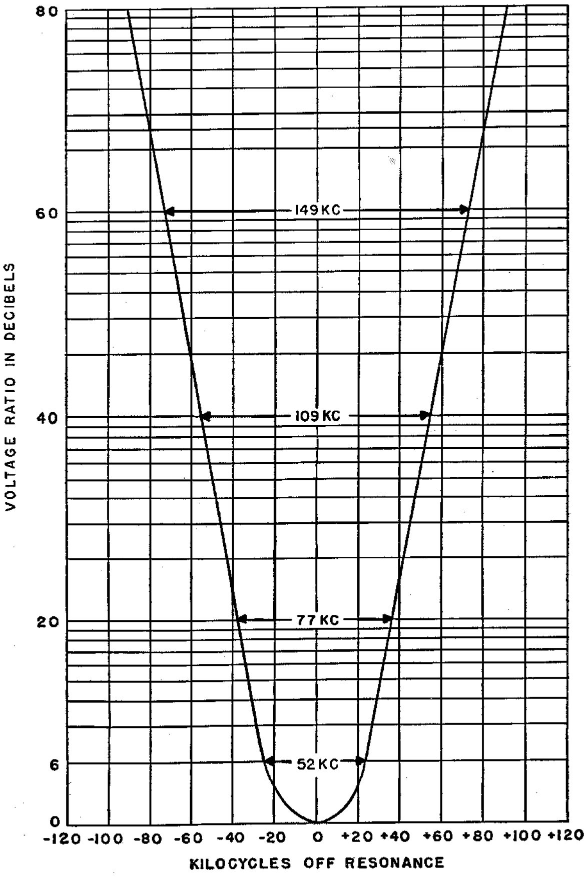

Fig. 4—Intermediate-frequency selectivity characteristic. Possible future assignment of 100-kilocycle channels in the band requires a steep slope at the sides of the curve. Center frequency is 6.325 megacycles.

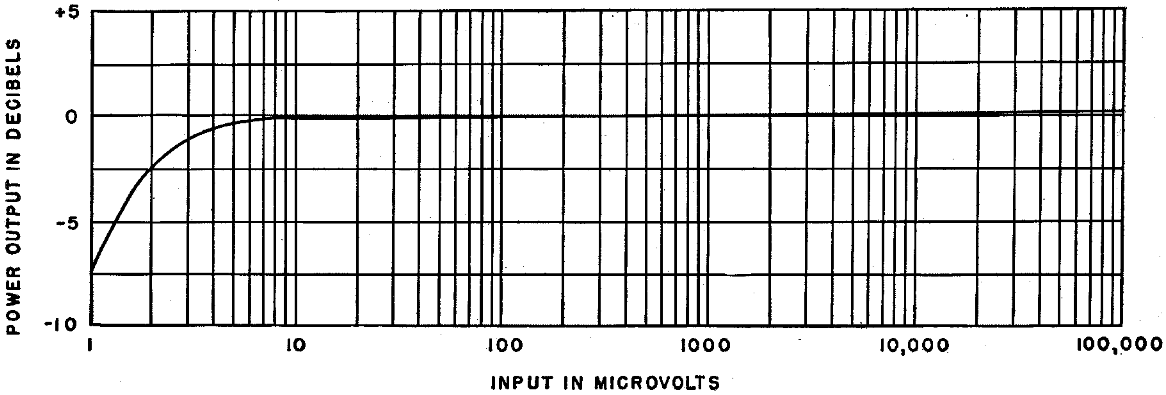

Fig. 5—Automatic-volume-control characteristic. Because of the generally low power of aircraft transmitters, and their widely variable distance from the receiver, the design must prevent loss of weak signals and blocking on the stronger ones. Zero level is 6 milliwatts in 600 ohms.

maximum receiver sensitivity prevails below the 2-microvolt level. This characteristic permits relatively weak signals to deliver a zero-level audio-frequency signal into a telephone line for remote operation. This feature is intended to preclude any need for remote radio-frequency gain control, although provision for that control is incorporated to provide for a remote squelch control.

2.5 Noise Limiter and Squelch

The noise-limiter circuit employed in this receiver is a very effective means of suppressing ignition noise, static, and other forms of recurrent pulses without distorting the audio-frequency signal. Fig. 6 shows the circuit, including the squelch, in schematic form. The theory of operation is explained as follows.

Under steady carrier conditions and with the squelch switch in the off position, the plate voltage of *V1* is more positive than that of *V2*, and therefore diodes *V3* and *V4* conduct. The relatively low impedance of diode *V4* allows the audio-frequency voltage to pass to the amplifier. Upon the application of a sharp noise pulse, the plate voltage of *V2* increases, but the plate voltage on *V1* remains momentarily constant due to the time constant of *R5* and *C2* in the grid circuit of *V1*. When the plate voltage of *V2* exceeds the plate voltage of *V1*, diodes *V3* and *V4* block, and the audio-frequency voltage is cut off for the duration of the pulse. An accelerating action, which tends to speed the blocking of the diodes, results from *C1* and *R6*.

The rectified noise pulse develops a negative voltage at the plate of *V4* and a more positive voltage on the cathode of *V3*, so that a fast blocking is obtained. When this occurs, the audio-frequency voltage at that instant will be held constant across *C3*, as there is no path for discharge. When the diodes again conduct, due to the decay of the noise pulse, the voltage across *C3* will then follow the instantaneous level. This provides a continuous audio-frequency signal with very little noise.

Voltage divider *R2* permits adjustment of the clipping level for modulation peaks from 60 to 100 percent. Although this control is normally set to pass a 100-percent-modulated wave, it is sometimes desirable in excessively noisy locations to improve the effectiveness of the noise limiter by reducing the bias on the grid of *V1*. This will tend to introduce distortion on highly modulated signals, but an improved signal-to-noise ratio will result and may be the more desirable.

By the addition of resistor *R10* in the plate circuit of *V1*, an inherent squelch action is obtained; with no signal input, the plate voltage of *V1* is lower than that of *V2* (because of the different values of plate-load resistance), and the diodes do not conduct. Upon arrival of a carrier of predetermined strength, dependent upon the amount of gain, the bias on *V1* becomes higher than the bias on *V2*, causing more voltage to appear at the plate of *V4* than at the cathode of *V3*, and the diodes conduct.

The radio-frequency gain control incorporated in the receiver allows control of the squelch for signal inputs ranging from 1.5 to 1000 microvolts. This is an advantage to control-tower operators who want a muted receiver at noisy sites when no signal is on their frequency; they can adjust the gain control to bring a given noise level to the threshold of the squelch and still be confident of receiving weak signals.

2.6 Output Circuit

The audio-frequency amplifier system is capable of delivering a maximum of one watt into a 600-ohm output impedance. The audio-frequency fidelity is designed for speech operation and is flat to within ±3 decibels from 350 to 3500 cycles per second. The output transformer has a split-and-balanced secondary to provide maximum flexibility of load connections.

By simplexing the grid bias of the five gain-controlled tubes on the output circuit, the radio-frequency gain and squelch may be controlled remotely over a telephone line. *Simplexing* is a term meaning a system whereby a single 2-wire physical circuit may be made to carry independent direct-current and alternating-current signals; separation of these components being accomplished at the terminals by proper arrangement of inductances and capacitances.

A monitoring phone jack is connected across one section of the output-transformer secondary in order that the monitoring may be accomplished at the local installation independently of all external connections.

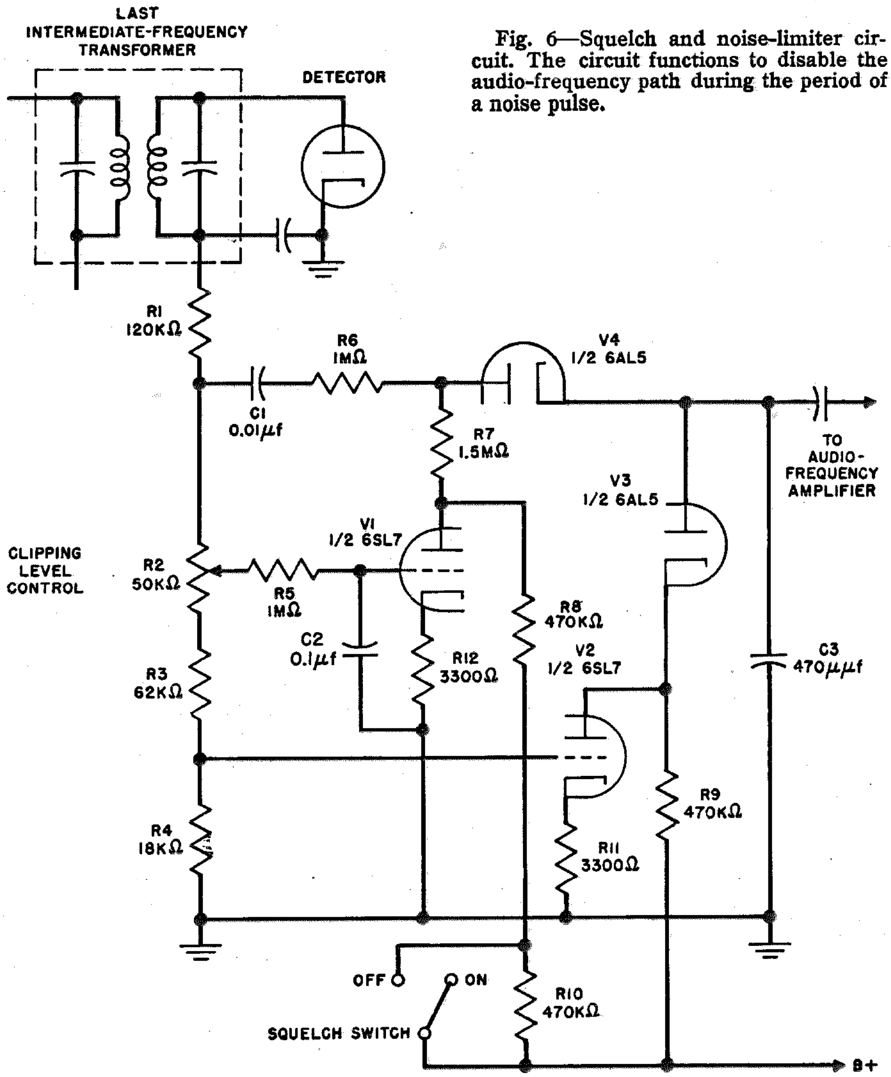

Fig. 6—Squelch and noise-limiter circuit. The circuit functions to disable the audio-frequency path during the period of a noise pulse.

3. *Field Test*

The receiver has been installed and operated successfully in many control towers of major airports. In the Chicago airport control tower, sensitivity checks were made with Civil Aeronautics Administration cooperation. An outgoing plane flying at 4000 feet called the tower every five minutes. Due to heavy construction work in progress at the airport, a 5-microvolt noise level was induced in the antenna. The noise limiter incorporated in the receiver was so effective that signals from the plane's 5-watt transmitter were received from a distance of more than 120 miles with a good signal-to-noise ratio.

In order to determine, under practical operating conditions, the effectiveness of the receiver selectivity in rejecting signals 100 kilocycles off resonance, field tests were arranged with the cooperation of American Airlines Incorporated. An antenna installed on the Lincoln Building (750 feet high) in Manhattan, was connected by about 50 feet of Type RG-8/U cable to the receiver, which was tuned to 126.7 megacycles. Carriers received from LaGuardia Airport (8 miles away), where a 50-watt transmitter was operating on 126.7 megacycles, developed a signal of approximately 1200 microvolts at the receiver input terminals. An identical receiver, adjusted for 126.8 megacycles, was then substituted. Due to the high order of adjacent-channel attenuation, no signal was heard on this receiver even though the squelch circuit was rendered inoperative and the receiver gain was at maximum.

Using the same receiving-antenna installation, the test was repeated under more severe conditions by using a 50-watt transmitter (126.7 megacycles) installed in the Lincoln Building. The spacing between the transmitting and receiving antennas was about 100 feet. This transmitter developed over 0.2 volt at the receiver input terminals. No blocking of the 126.7-megacycle receiver was noted. When the 126.8-megacycle receiver was substituted, an output signal equivalent to that produced by a 70-microvolt signal at resonance was obtained. Hence, the adjacent-channel selectivity attenuated the 126.7-megacycle 0.2-volt signal by more than 70 decibels. These test results therefore confirmed the laboratory measurements, and demonstrated the effectiveness of the radio- and intermediate-frequency amplifier design.

High Gain With Discone Antennas*

By A. G. KANDOIAN, W. SICHAK, and R. A. FELSENHELD

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EXTENSION of work on discone-type antennas, originally presented before the Institute of Radio Engineers in 1945, is described. Since that time, these antennas have been developed for a large number of applications. Of particular interest are those applications that require high gain over a very broad band of frequencies. Detailed data are given on several designs, particularly on unidirectional antennas for point-to-point communication and omnidirectional high-gain antennas for the distance-measuring equipment (DME) of the Navar series of aids to aerial navigation.

• • •

At frequencies above about 30 megacycles per second, it is possible to produce, conveniently and economically, a great deal of directivity of radiation. For some services, high-gain antennas must also provide omnidirectional radiation in the horizontal plane. The amount of antenna power gain obtainable is then limited because concentration of only the vertical radiation is permissable.

Such antennas have been developed and widely used in very-high-frequency broadcasting and in radar beacons in the ultra- and super-high-frequency spectra. As a rule, however, the bandwidth requirements have been relatively small. For most applications, horizontal polarization has been used.

1. *Omnidirectional High-Gain Broad-Band Antenna*

The present antenna development was undertaken for the ground station of the Navar distance-measuring equipment (DME). Among the requirements specified were a frequency range of 960 to 1215 megacycles, vertical polarization, omnidirectional horizontal radiation pattern, power gain of approximately 8 over a half-wave dipole, and a beam direction tilted up approximately one-half beam width with respect to the horizon over the complete frequency range. There were also mechanical and service requirements common to such applications.

The basic radiating unit selected to meet these requirements is the discone antenna,[1] which in itself is an omnidirectional vertically polarized radiator with broad-band characteristics well in excess of the present requirements. By stacking a number of discones, one above the other, and supplying them with equal power in proper phase, the vertical pattern may be compressed and, hence, the over-all antenna gain greatly increased. An array of this type of nine discones is shown in Fig. 1.

The radiation pattern of such an array is given by

$$F(\beta) \cong \cos\beta \, \frac{\sin n\left(\dfrac{S^\circ}{2}\sin\beta - \dfrac{\phi}{2}\right)}{\sin\left(\dfrac{S^\circ}{2}\sin\beta - \dfrac{\phi}{2}\right)}, \qquad (1)$$

where β = vertical angle, zero at the horizon,

n = number of discones,

S° = spacing between successive radiators in electrical degrees, and

ϕ = phase difference between successive radiators.

When all the radiators are operated in phase, ϕ is zero and $F(\beta)$ is maximum for $\beta = 0$; that is, the radiation is always maximum in the horizontal direction. Fig. 2 shows a calculated pattern of such an array of 9 stacked elements. The power gain of an array of stacked discones is essentially the same as for an array of stacked horizontal loop antennas as given in reference 1.

Energizing a number of radiators in phase can be accomplished in many ways for a fixed frequency or a limited frequency band. Over a large frequency range, this may only be accomplished by making all the transmission lines to the individual radiators of equal length. This makes the relative phase of the currents in the various radiators equal and independent of frequency.

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* Presented, National Electronics Conference, Chicago, Illinois, November 4, 1947. Reprinted from *Proceedings of the National Electronics Conference*, v. 3, pp. 336–346; 1947: figures 12, 13, and 14 were not included in the original publication.

[1] A. G. Kandoian, "Three New Antenna Types and Their Application," *Proceedings of the I.R.E.*, v. 34, pp. 70W–75W; February, 1946: Also, *Electrical Communication*, v. 23, pp. 27–34; March, 1946.

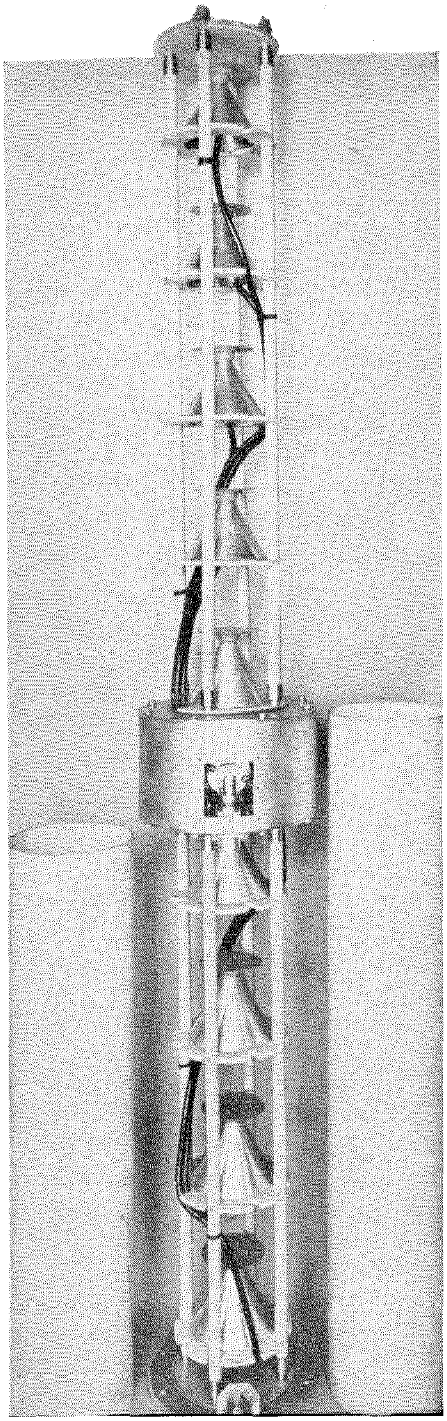

Fig. 1—Discone array used for the distance-indicator function of the Navar system. Consisting of 9 discone radiators stacked vertically, this omnidirectional antenna operates anywhere within the range of 960 to 1215 megacycles. The mountings and supporting rods are constructed of fibre-glass, as are the covers standing at the sides of the array.

In the present case, it is desired to tilt the pattern approximately one-half beam width above the horizon throughout the frequency range. This may be accomplished by supplying equal power to all the radiators as before, but successively delaying the phase of the current in each element from the bottom to the top. This relative phase is represented by ϕ in (1). The value of ϕ for a given tilt of pattern is

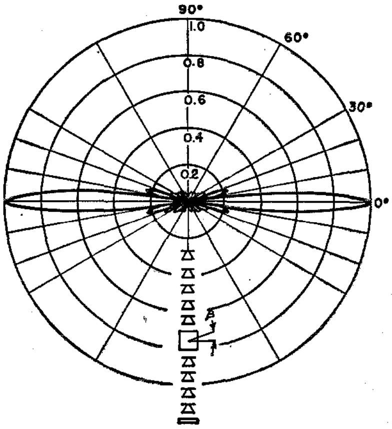

Fig. 2—Calculated vertical radiation pattern of a stack of 9 discones all energized in phase;

$$F(\beta) \cong \cos\beta \frac{\sin[n(140 \sin\beta)]}{\sin[140 \sin\beta]}.$$

$$\phi = S^{\circ} \sin \beta_T, \qquad (2)$$

where β_T is the tilt of the beam with respect to the plane perpendicular to the antenna axis. This is derived simply by maximizing (1).

Thus, to produce a tilt in the beam direction of 3.5 degrees and for spacing S° equal to 280 degrees between successive discones, a phase stagger of $\phi = 15$ degrees is obtained. Fig. 3 shows the calculated pattern of such an array of 9 discones.

This establishes the basic theory of a high-gain omnidirectional antenna whose beam direction may be shifted electrically.

To reduce this to practice, first let us consider the simpler problem of an omnidirectional broadside antenna without beam tilting. The major problem is to divide the available power equally among a large number of elements and keep the currents in phase over a wide frequency range. This is accomplished in the present case by making the input impedance of each radiating element equal to the surge impedance of some standard RG-type cable. The discones used in this array meet this requirement very well in that when used with RG-8/U cable they have a mismatch of less than 1.2 over a frequency range much larger than 960 to 1215 megacycles. The individual feeders are all connected in parallel at a common junction box and then transformed back to 50 ohms impedance by a simple quarter-wave transformer. The bandwidth limitation of the array is now determined by the quarter-wave transformer, rather than the radiators themselves and, wherever justified, larger bandwidths could be obtained by multiple sections or exponential transformers.

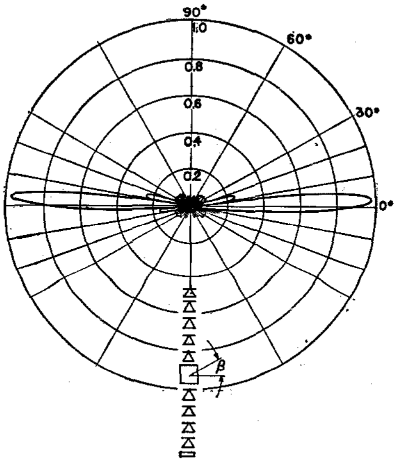

Fig. 3—Calculated vertical radiation pattern for array of Fig. 1 operating at 1087 megacycles;

$$F(\beta) \cong \cos\beta \frac{\sin n\left(\frac{S^\circ}{2}\sin\beta - \frac{\phi}{2}\right)}{\sin\left(\frac{S^\circ}{2}\sin\beta - \frac{\phi}{2}\right)}, \quad \text{and} \quad \phi = S^\circ \sin\beta_T.$$

The next step, that of tilting the beam above the horizon, is now very straightforward. All it is necessary to do is to make successive lengths of RG-8/U cable leading from the common junction box to the individual discones such that each one is 15 degrees longer than the one below it. This amounts to a length differential of only three-tenths of an inch between successive feeders.

There are certain possible complications in an array of this type, which tend to alter the input impedance and the radiation pattern. The major problem is the placement of the individual feeders. It was determined experimentally that least

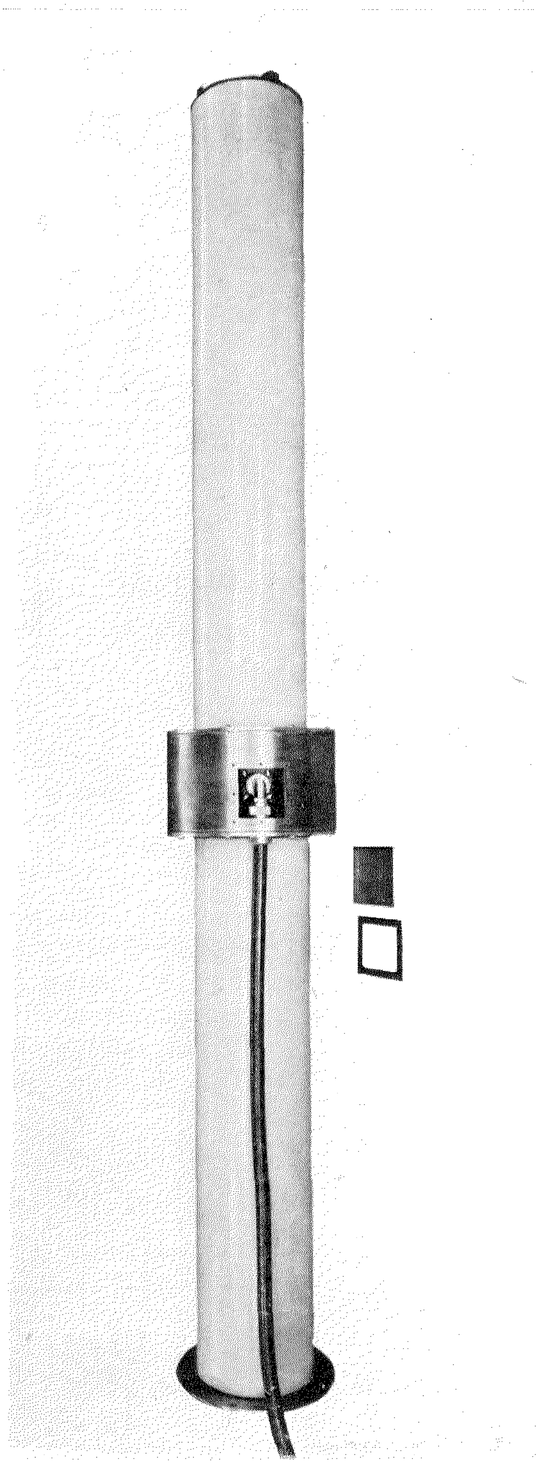

Fig. 4—Complete 9-element discone array with weatherproofing covers mounted. The access plate has been removed from the cable-connector housing, showing the type-N connector.

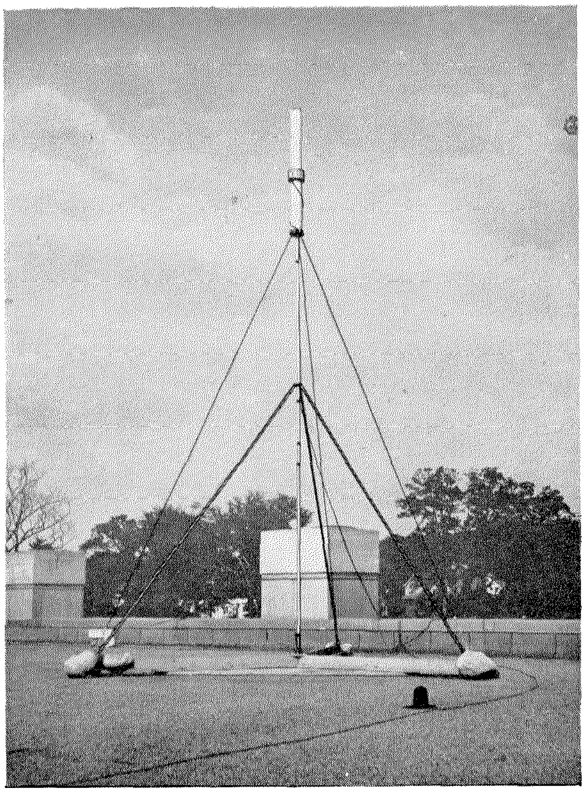

Fig. 5—Array mounted on mast on laboratory roof for flight tests with distance-measuring equipment.

disturbance was noticed when the cables were strapped together and then wound in a spiral around the discones, starting from the center of the array and working toward the top and bottom. The bunching gives the least reflection and the spiral tends to minimize the pattern distortion by spreading the effects evenly around the antenna. The excess length of cable feeding the center radiators is coiled up inside the metallic housing near the center of the array.

Mechanically, the antenna array is mounted on a fibre-glass supporting structure and enclosed from the weather by a fibre-glass housing. This results in a very strong and electrically efficient design. Fig. 4 shows the antenna unit complete in its housing. The over-all height of the structure is approximately 72 inches. Fig. 5 shows the array, 50 feet above ground level, installed in actual operation on top of a mast.

Measured data on the antenna array throughout the frequency range are given in Figs. 6 to 10. It is to be noted that though the half-power beam width narrows from 8 to 6 degrees from 960 to 1215 megacycles, the beam direction remains fixed at approximately 3.5 degrees above the horizon. The measured power gain was in all cases within ±1 decibel of the calculated[1] value.

2. Unidirectional Pencil-Beam Antenna

In considering the problem of producing a unidirectional pencil-beam antenna as free as possible of any frequency limitations, the combining of a paraboloid and a discone antenna is a logical step. Measurements have shown that such a combination produces beam patterns that are satisfactory for communication and other applications over an extremely wide fre-

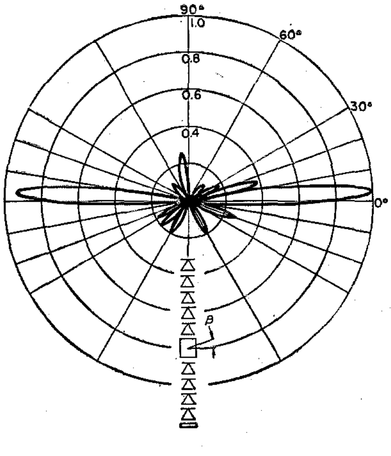

Fig. 6—Measured vertical radiation pattern at 960 megacycles.

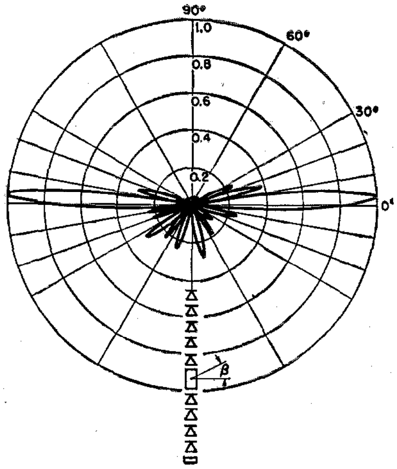

Fig. 7—Measured vertical radiation pattern at 1087 megacycles.

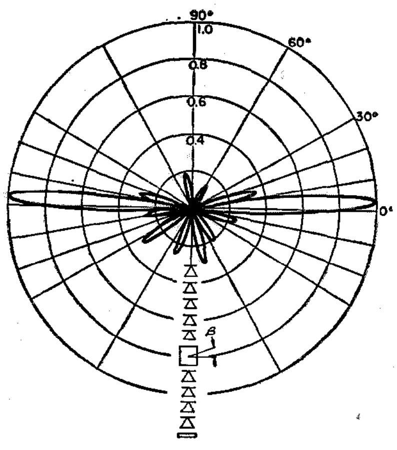

Fig. 8—Measured vertical radiation pattern at 1215 megacycles.

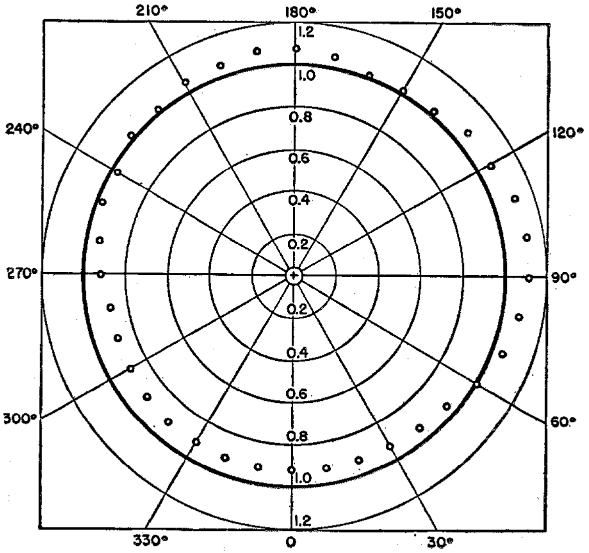

Fig. 9—Horizontal radiation pattern at 1087 megacycles. The calculated pattern is shown as a solid line and the measured values are indicated by small circles.

quency range. The standing-wave ratio is less than 2 over a 4.5-to-1 frequency range (700 to 3100 megacycles). As half of the power radiated by the discone is not intercepted by a focus-in-aperture paraboloid, it is to be expected that the gain will be somewhat lower than if a more directive source were used. If the paraboloid is 10 or more wavelengths in diameter, so that the effect of the direct radiation from the discone will be small compared with the total radiation from the paraboloid, the gain relative to an isotropic radiator is[2] approximately

$$G = 0.37\left(\frac{\pi D}{\lambda}\right)^2, \tag{3}$$

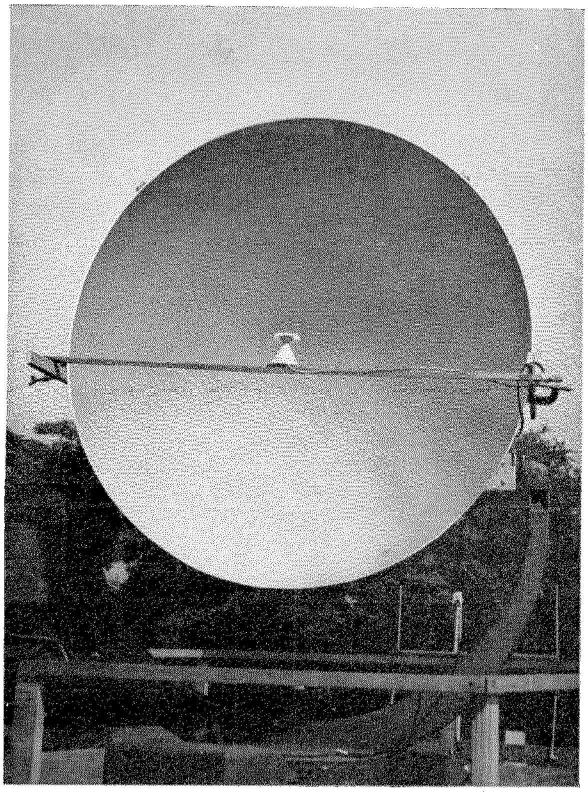

Fig. 11—A single discone radiator and 72-inch parabolic reflector. This arrangement is quite effective where an antenna for a pencil-shaped beam is required to have broad-bandwidth characteristics.

where D = diameter and
λ = wavelength.

This gain is about 4.5 decibels below that of a uniformly irradiated parabolic antenna, and about 2.5 decibels below that of a well-designed narrow-band antenna using a simple dipole and reflector. If the paraboloid is only a few wavelengths in diameter, so that the direct radiation

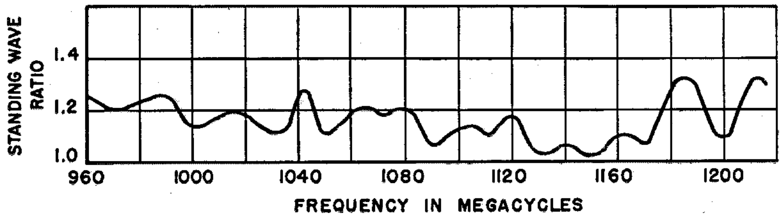

Fig. 10—Standing-wave ratio plotted against frequency. Measured through 50 feet of RG-14/U cable on a 50-ohm slotted line.

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[2] F. E. Terman, "Radio Engineers' Handbook," McGraw-Hill Book Company, New York, New York, 1943, pp. 839–840.

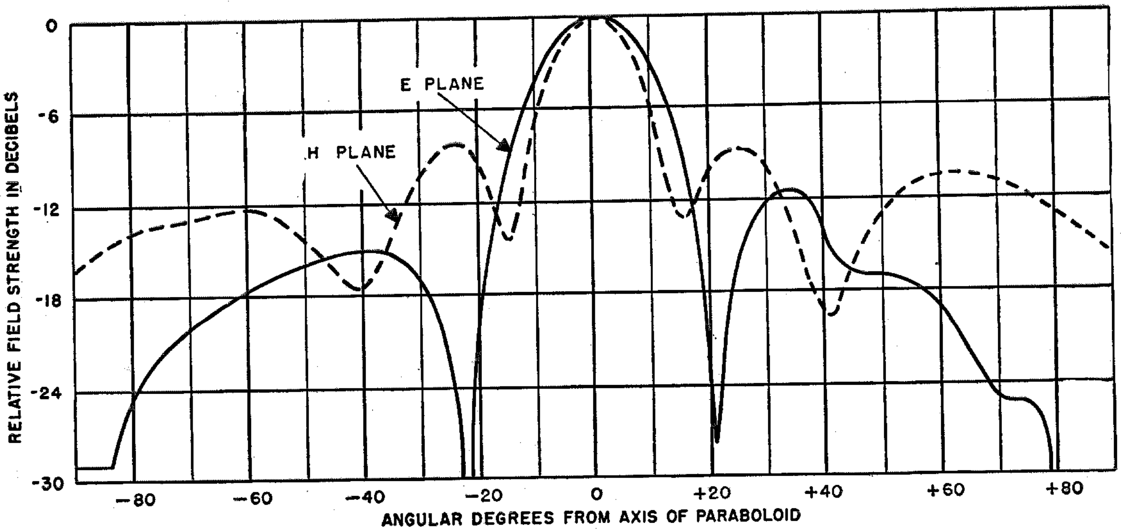

Fig. 12—The solid line indicates the *E*-plane (vertical), and the dashed line indicates the *H*-plane (horizontal) beam pattern of the antenna of Fig. 11. The measurements were made at 600 megacycles.

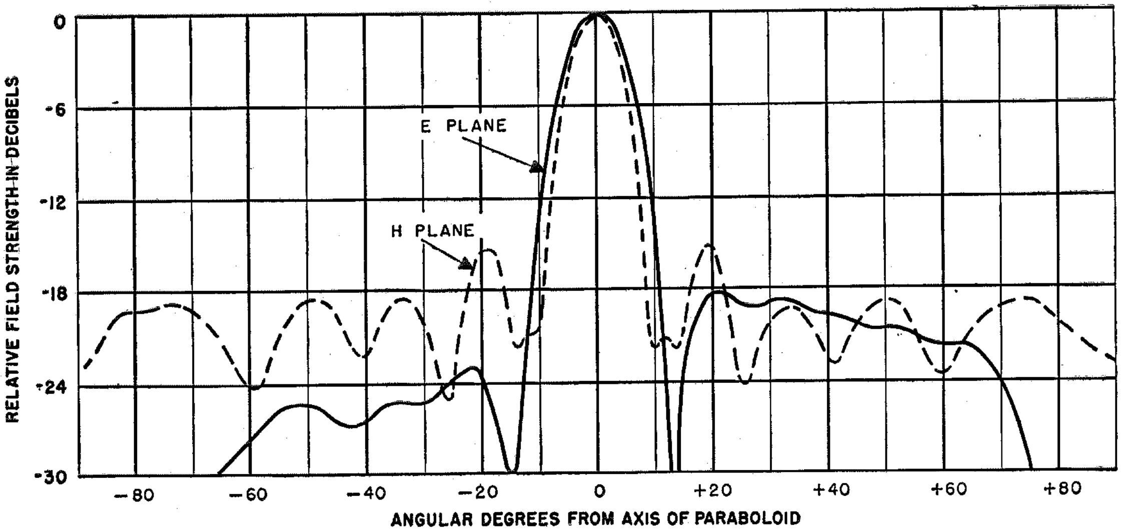

Fig. 13—Parabolic-antenna beam patterns measured at 1290 megacycles.

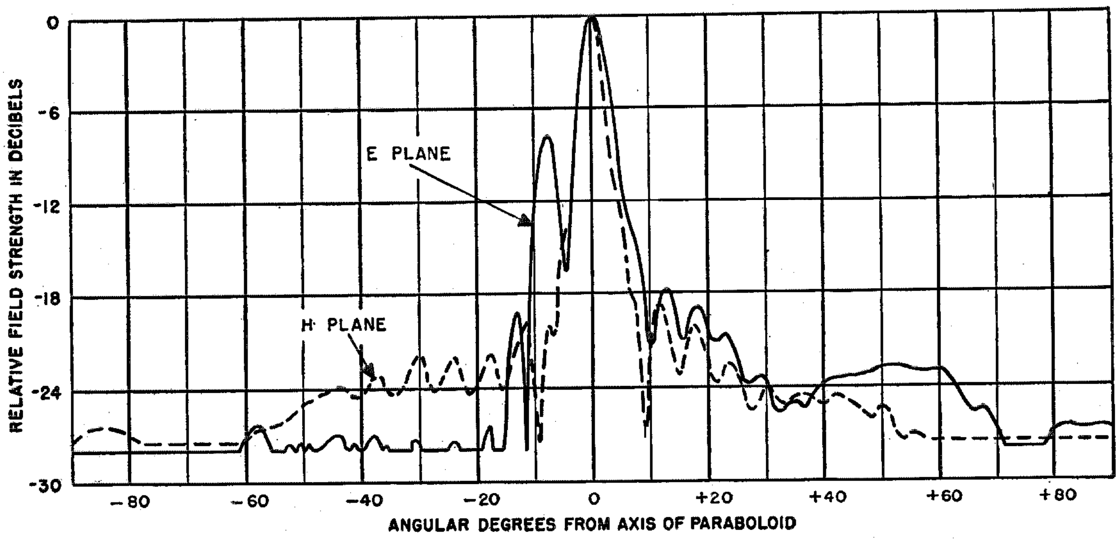

Fig. 14—Parabolic-antenna beam patterns measured at 3200 megacycles.

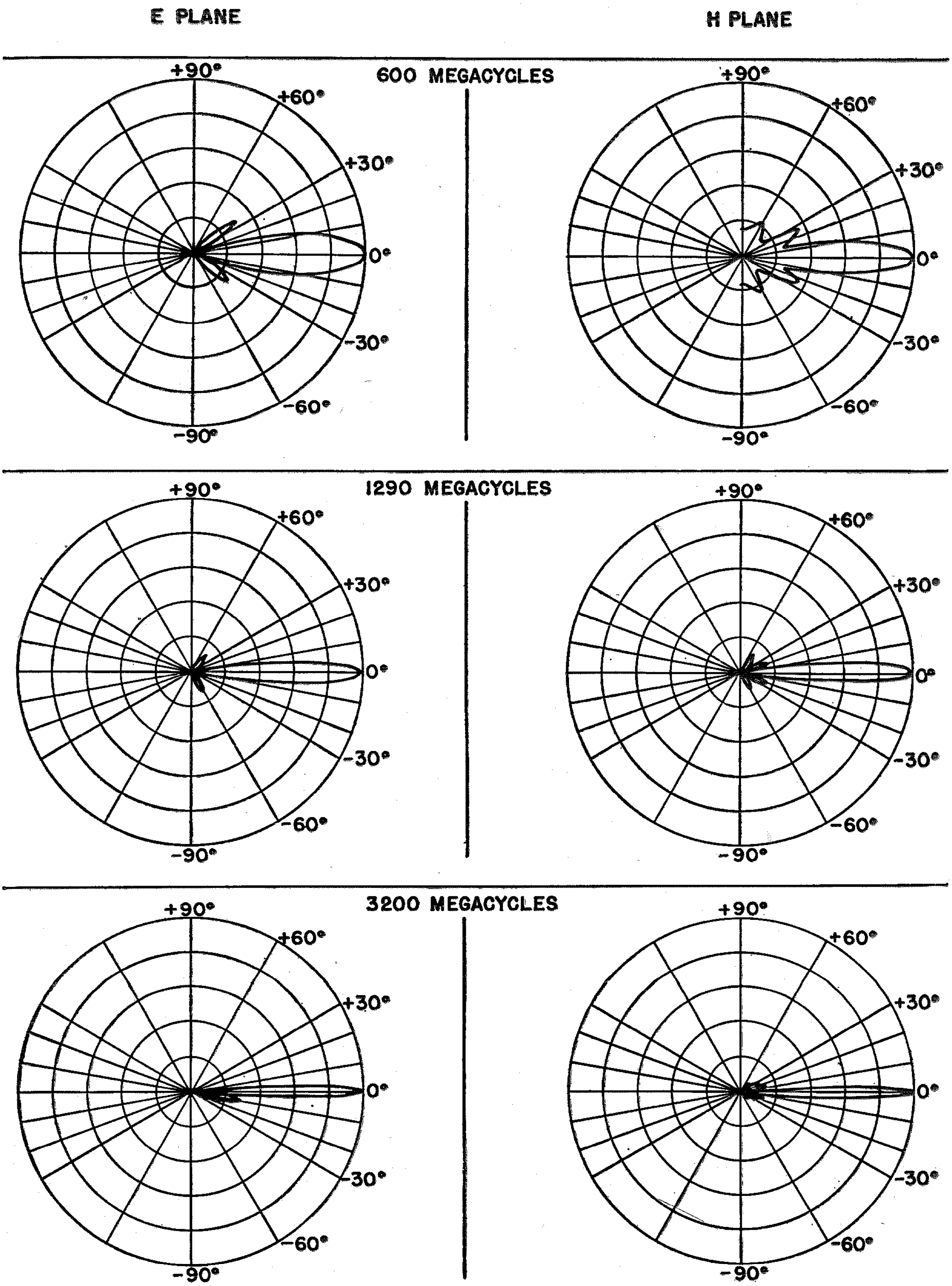

Fig. 15—*E*- and *H*-plane patterns of Figs. 12, 13, and 14 plotted on polar coordinates.

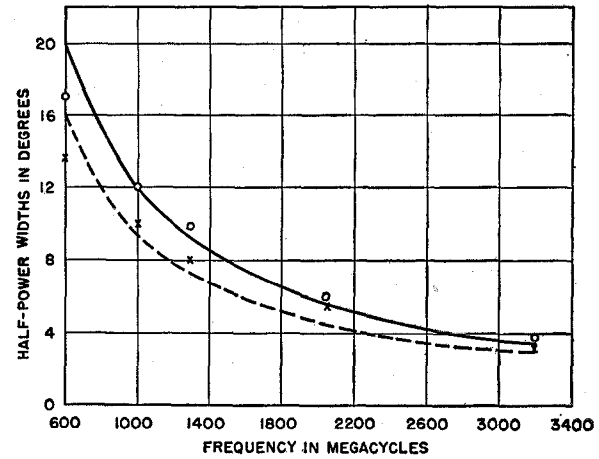

Fig. 16—Variation of the half-power beam widths with frequency for the antenna of Fig. 11. The circles indicate measured results for the *E*-plane pattern; the small crosses, the results for the *H* plane. The solid line indicates the theoretical widths for the *E* plane, calculated from $\theta_B = 1.27\lambda/D$. The dashed line indicates the *H*-plane widths derived from $\theta_B = \lambda/D$. θ_B is the half-power beam width.

becomes an appreciable portion of the total, the gain will depend on the relative phase between the direct radiation and the reflected radiation; the gain may then be larger or smaller than that given by (3). Equation (3) indicates that a 72-inch-diameter paraboloid will have a gain varying from 16.5 decibels at 700 megacycles to 29.5 decibels at 3200 megacycles.

The discone discussed above was used as a radiation source in a 72-inch-diameter paraboloid having a 21.1-inch focal length. The discone was supported by a piece of wood $\frac{3}{4}$ inch wide by $1\frac{1}{4}$ inches thick, which was fastened across the face of the paraboloid (Fig. 11). The discone itself, like the ones used in the distance-measuring-equipment array, consists of a 60-degree cone $4\frac{1}{4}$ inches high and a $3\frac{3}{4}$-inch disc supported above the cone and separated from it by polyethylene insulation. A standard type-N connector is located inside the cone. The RG-8/U cable, which ran from one edge of the paraboloid to the discone, was attached to the wooden support.

Principal electric-plane and magnetic-plane patterns were measured at frequencies of 600, 1000, 1290, 2040, and 3200 megacycles. Some of these patterns are shown in Figs. 12, 13, and 14. In Fig. 15, these same patterns are plotted, more conventionally, in plan coordinates. The measurements were made with the receiving antenna 70 feet from the paraboloid. Fig. 16 shows both a theoretical and measured plot of half-power beam width with frequency. The theoretical values were obtained from patterns calculated on the basis of physical optics.

The standing-wave ratio was measured in the standard manner with 10 feet of RG-8/U cable between the standing-wave detector and the discone. A 6-inch tapered section was used between the standing-wave detector and the cable fitting. Between 850 and 3200 megacycles, the frequency was varied in steps of approximately 20 or 30 megacycles. Four readings were taken between 520 and 680 megacycles; no readings were taken between 680 and 850 megacycles. The standing-wave ratio is plotted against frequency in Fig. 17. Also given in Fig. 17 is the standing-wave ratio of the discone in free space.

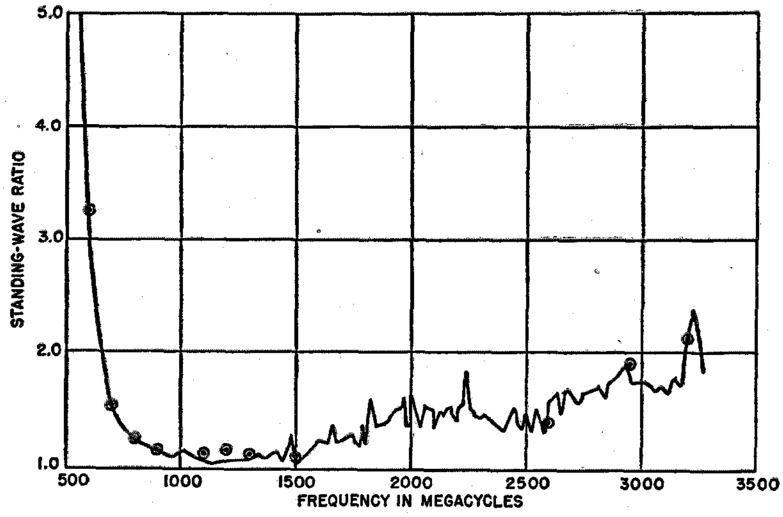

Fig. 17—Standing-wave ratio plotted against frequency for the antenna of Fig. 11. The irregularities are due to the effects of the coaxial cable and associated fittings. Measured results for a discone radiator alone in space are given by the circled points.

The rapid fluctuation in the standing-wave ratio is probably due to the type-N cable connectors and the variations in the dielectric cable itself. These effects are well known and will not be discussed here. The measured standing-wave ratio is slightly lower than that of the antenna itself, due to the cable attenuation. The nominal attenuation of 10 feet of RG-8/U cable is 0.85 decibel at 1000 megacycles, and 1.7 decibels at 3000 megacycles. At 1000 megacycles, a standing-wave ratio of 1.2 will, therefore, be transformed to 1.17, while at 3000 megacycles, a standing-wave ratio of 2.0 will be transformed to 1.74. But in actual operation, some cable must be used, so that it is the standing-wave ratio of the combination that is of interest.

If the focal length of the paraboloid is more than a few wavelengths, the voltage-reflection coefficient introduced into the transmission line by the paraboloid is given[3] by

$$\Gamma = \frac{G\lambda}{4\pi F}, \tag{3}$$

where G = gain of feed in direction of vertex,
λ = wavelength, and
F = focal length.

The gain of the discone being substantially the same as that of a half-wave dipole, this formula gives a reflection coefficient of 0.075 (standing-wave ratio of 1.16) at 1000 megacycles, and a reflection coefficient of 0.025 (standing-wave ratio of 1.05) at 3000 megacycles. This reflection will add to, or subtract from, the reflection of the discone alone, depending on the relative phase between the two reflections. If the reflections add in phase at 2000 megacycles, they will again add in phase at 1750 megacycles, and at 2270 megacycles, assuming that the phase of the discone reflection is constant at these frequencies. This effect is not evident in Fig. 17, but appears to be masked by the other reflections present.

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[3] S. Silver, Report No. 442, Radiation Laboratory, Massachusetts Institute of Technology.

Modern Demonstration of MacDonald's Equivalance Theorem

By A. G. CLAVIER

Federal Telecommunication Laboratories, Incorporated, Nutley, New Jersey

A DEMONSTRATION was given in 1912, by MacDonald,[1] of the following proposition, generally known as the "Equivalence Theorem":

> The electromagnetic field at any point P outside of a closed surface S due to the movement of electric charges inside S, can be expressed in terms of the electric and magnetic fields produced on S by said electric charges.

MacDonald did not make use of vectorial theory, so that his demonstration was rather lengthy. His method has been followed in the present paper, but considerably shortened by means of suitable notations. An interesting consequence is that it leads directly to the formulas given by Kottler[2] in the case of a harmonic time factor for the contribution of any portion of surface S limited by a closed contour C. It is not necessary to pass through the intermediary of fictitious electric- and magnetic-current densities on surface S.

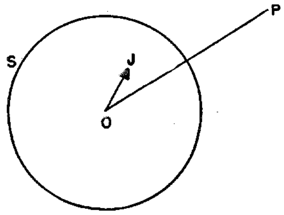

Fig. 1.

Let moving charges be situated inside a closed surface S (Fig. 1), the whole space being homogeneous and isotropic, with inductive capacitances ϵ and μ.

At all points of such a system, the field vectors $\boldsymbol{E}$ and $\boldsymbol{H}$ satisfy the following equations:

$$\operatorname{curl} \boldsymbol{E}+\mu\frac{\partial \boldsymbol{H}}{\partial t}=0, \qquad \operatorname{div} \boldsymbol{H}=0,$$

$$\operatorname{curl} \boldsymbol{H}-\epsilon\frac{\partial \boldsymbol{E}}{\partial t}=J, \qquad \operatorname{div} \boldsymbol{E}=\frac{\rho}{\epsilon}.$$

Factors ρ and J are the charge and current densities at the point considered. They are assumed to be zero for all points situated outside S.

The field vectors at point P, excluding electric fields not varying with time, can be expressed by means of the retarded-vector potential

$$\boldsymbol{A}_P=\frac{\mu}{4\pi}\int_V \frac{1}{r}\, J^* dv,$$

where V is the volume of S. The symbol J^* means that the value of $\boldsymbol{A}$ at time t depends on the value of J at time $t-r/c$ where c is the speed of light in the unbounded medium characterized by the factors ϵ and μ.

Now, r is the distance OP between the point O, where J is applied, and point P at which the field vectors are considered. These vectors satisfy the system

$$\boldsymbol{H}_P=\frac{1}{\mu}\operatorname{curl}_P \boldsymbol{A}_P,$$

$$\epsilon\frac{\partial \boldsymbol{E}_P}{\partial t}=\operatorname{curl}_P \boldsymbol{H}_P=\frac{1}{\mu}\operatorname{curl}_P\cdot\operatorname{curl}_P \boldsymbol{A}_P,$$

where curl_P means that the curl operator implies the coordinates of point P.

The first step of the demonstration consists of expressing the fields at P in terms of the fields at all points O where J is not zero;

$$\boldsymbol{H}_P=\frac{1}{4\pi}\operatorname{curl}_P\int_V \frac{1}{r}\, J^* dv,$$

$$\frac{J^*}{r}=\frac{(\operatorname{curl}_O \boldsymbol{H}_O)^*}{r}-\epsilon\frac{\partial}{\partial t}\frac{\boldsymbol{E}_O^*}{r},$$

$$\frac{J^*}{r}=\operatorname{curl}_O\frac{\boldsymbol{H}_O^*}{r}+\operatorname{curl}_P\frac{\boldsymbol{H}_O^*}{r}-\epsilon\frac{\partial}{\partial t}\frac{\boldsymbol{E}_O^*}{r}.$$

From these equations, the value of $\mu(\partial H_P/\partial t)$ can be derived. Note that

$$\mu\frac{\partial \boldsymbol{H}_O}{\partial t}=-\operatorname{curl} \boldsymbol{E}_O,$$

therefore,

$$\frac{\mu\frac{\partial \boldsymbol{H}_O}{\partial t}}{r}=-\frac{(\operatorname{curl} \boldsymbol{E}_O)^*}{r}=-\operatorname{curl}_O\frac{\boldsymbol{E}_O^*}{r}-\operatorname{curl}_P\frac{\boldsymbol{E}_O^*}{r},$$

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[1] MacDonald, *Proceedings of the London Mathematical Society*, v. 10, p. 91; 1912.

[2] Kottler, *Annalen der Physik*, v. 71, p. 457; 1923.

so that

$$\mu\frac{\partial \boldsymbol{H}_P}{\partial t}=\frac{\partial}{\partial t}\operatorname{curl}_P\frac{\mu}{4\pi}\int_V \operatorname{curl}_O\frac{\boldsymbol{H}_O^*}{r}dv$$
$$-\operatorname{curl}_P\cdot\operatorname{curl}_P\frac{1}{4\pi}\int_V\operatorname{curl}_O\frac{\boldsymbol{E}_O^*}{r}dv$$
$$-\frac{1}{4\pi}\operatorname{curl}_P\int_V\left(\operatorname{curl}_P\cdot\operatorname{curl}_P\frac{\boldsymbol{E}_O^*}{r}+\epsilon\mu\frac{\partial^2}{\partial t^2}\frac{\boldsymbol{E}_O^*}{r}\right)dv.$$

The third part of the left-hand side of this equation is, however, equal to zero. This is shown by rewriting

$$-\frac{1}{4\pi}\int_V\operatorname{curl}_P\left(\operatorname{grad}_P\operatorname{div}_P-\Delta_P+\frac{1}{c^2}\frac{\partial^2}{\partial t^2}\right)\frac{\boldsymbol{E}_O^*}{r}dv$$
$$=\frac{1}{4\pi}\operatorname{curl}_P\int_V\left(\Delta_P-\frac{1}{c^2}\frac{\partial^2}{\partial t^2}\right)\frac{\boldsymbol{E}_O^*}{r}dv.$$

Each component of this vector is easily found to be zero.

Let

$$\boldsymbol{A}_v=\frac{\mu}{4\pi}\int_V\operatorname{curl}_O\frac{\boldsymbol{H}_O^*}{r}dv$$

and

$$\boldsymbol{A}_v'=\frac{-\epsilon}{4\pi}\int_V\operatorname{curl}_O\frac{\boldsymbol{E}_O^*}{r}dv,$$

then,

$$\frac{\partial \boldsymbol{H}_P}{\partial t}=\frac{1}{\mu}\operatorname{curl}_P\frac{\partial \boldsymbol{A}_v}{\partial t}+c^2\operatorname{curl}_P\cdot\operatorname{curl}_P\boldsymbol{A}_v',$$

and disregarding any static distribution of $\boldsymbol{H}_P$ and $\boldsymbol{E}_P$;

$$\frac{\partial \boldsymbol{E}_P}{\partial t}=-\frac{1}{\epsilon}\operatorname{curl}_P\frac{\partial \boldsymbol{A}_v'}{\partial t}+c^2\operatorname{curl}_P\cdot\operatorname{curl}_P\boldsymbol{A}_v.$$

It is thus seen tht the variable fields $\boldsymbol{E}_P$ and $\boldsymbol{H}_P$ at point P can be expressed in terms of the fields inside S by means of the two vectors $\boldsymbol{A}_v$ and $\boldsymbol{A}_v'$. It is now a simple matter to transform the volume integrals into surface integrals;

$$\int_V\operatorname{curl}_O\frac{\boldsymbol{H}_O^*}{r}dv=\int_S\left(\boldsymbol{N}^O\times\frac{\boldsymbol{H}_O^*}{r}\right)ds.$$

Let

$$\boldsymbol{A}_S=\frac{\mu}{4\pi}\int_S\left(\boldsymbol{N}^O\times\frac{\boldsymbol{H}_O^*}{r}\right)ds$$

and

$$\boldsymbol{A}_S'=-\frac{\epsilon}{4\pi}\int_S\left(\boldsymbol{N}^O\times\frac{\boldsymbol{E}_O^*}{r}\right)ds,$$

then,

$$\frac{\partial \boldsymbol{H}_P}{\partial t}=\frac{1}{\mu}\operatorname{curl}_P\frac{\partial \boldsymbol{A}_S}{\partial t}+c^2\operatorname{curl}_P\cdot\operatorname{curl}_P\boldsymbol{A}_S',$$
$$\frac{\partial \boldsymbol{E}_P}{\partial t}=-\frac{1}{\epsilon}\operatorname{curl}_P\frac{\partial \boldsymbol{A}_S'}{\partial t}+c^2\operatorname{curl}_P\cdot\operatorname{curl}_P\boldsymbol{A}_S.$$

The computation of the field vectors at point P can obviously be made in two different ways: The first starts with the determination of vectors $\boldsymbol{A}_S$ and $\boldsymbol{A}_S'$. When this is done, the operations curl_P and $\operatorname{curl}_P\cdot\operatorname{curl}_P$ are performed; these operations involve derivatives with respect to the coordinates of point P only.

A second manner consists of interchanging the processes of integration and differentiation, writing

$$\frac{\partial \boldsymbol{H}_P}{\partial t}=\frac{1}{\mu}\frac{\partial}{\partial t}\frac{\mu}{4\pi}\int_S\operatorname{curl}_P\left(\boldsymbol{N}^O\times\frac{\boldsymbol{H}_O^*}{r}\right)ds$$
$$+c^2\frac{\epsilon}{4\pi}\int_S\operatorname{curl}_P\cdot\operatorname{curl}_P\left(\boldsymbol{N}^O\times\frac{\boldsymbol{E}_O^*}{r}\right)ds,$$

and the corresponding expression for $\partial\boldsymbol{E}_P/\partial t$. This is the same as attributing

$$\frac{1}{4\pi}\left[\frac{\partial}{\partial t}\operatorname{curl}_P\left(\boldsymbol{N}^O\times\frac{\boldsymbol{H}_O^*}{r}\right)\right.$$
$$\left.+\frac{1}{\mu}\operatorname{curl}_P\cdot\operatorname{curl}_P\left(\boldsymbol{N}^O\times\frac{\boldsymbol{E}_O^*}{r}\right)\right]$$

as the contribution of ds to the magnetic field, and

$$\frac{1}{4\pi}\left[-\frac{\partial}{\partial t}\operatorname{curl}_P\left(\boldsymbol{N}^O\times\frac{\boldsymbol{E}_O^*}{r}\right)\right.$$
$$\left.+\frac{1}{\epsilon}\operatorname{curl}_P\cdot\operatorname{curl}_P\left(\boldsymbol{N}^O\times\frac{\boldsymbol{H}_O^*}{r}\right)\right]$$

as the contribution of the same surface element to the electric field.

Now consider the operation $\operatorname{curl}_P\cdot\operatorname{curl}_P=\operatorname{grad}_P\cdot\operatorname{div}_P-\Delta_P$. The first term can be written as follows:

$$\operatorname{grad}_P\cdot\operatorname{div}_P\left(\boldsymbol{N}^O\times\frac{\boldsymbol{E}_O^*}{r}\right)ds$$
$$=\operatorname{grad}_P\frac{\boldsymbol{E}_O^*}{r}\cdot\operatorname{curl}\boldsymbol{N}^O\cdot ds-\operatorname{grad}\cdot\boldsymbol{N}^O\cdot\operatorname{curl}_P\frac{\boldsymbol{E}_O^*}{r}ds.$$

Now, $\operatorname{curl}_P\boldsymbol{N}^O\cdot ds=\boldsymbol{dl}$, the last being the element of the contour integral along the closed contour limiting any portion of surface S.

When there are no line discontinuities of the fields on the closed surface S, the integral $\int_C\operatorname{grad}_P\frac{\boldsymbol{E}_O^*}{r}\cdot\boldsymbol{dl}$ obviously disappears. It must remain, however, in the correct expression of the contribution of any portion ΔS of surface S, which is thus found to be

$$\frac{\partial \boldsymbol{H}_P}{\partial t}=\frac{\partial}{\partial t}\frac{1}{4\pi}\int_{\Delta S}\text{curl}_P\left(\boldsymbol{N}^O\times\frac{\boldsymbol{H}_O^*}{r}\right)ds$$
$$-\frac{1}{\mu}\frac{1}{4\pi}\int_{\Delta S}\text{grad }\boldsymbol{N}^O\cdot\text{curl}_P\frac{\boldsymbol{E}_O^*}{r}\cdot ds$$
$$-\epsilon\cdot\frac{1}{4\pi}\frac{\partial^2}{\partial t^2}\int_{\Delta S}\left(\boldsymbol{N}^O\times\frac{\boldsymbol{E}_O^*}{r}\right)ds+\frac{1}{4\pi}\int_C\frac{\boldsymbol{E}_O^*}{r}\cdot\boldsymbol{dl}.$$

A similar expression is readily obtained for $\partial \boldsymbol{E}_P/\partial t$. It is only necessary to interchange $\boldsymbol{E}_O^*$ and $\boldsymbol{H}_O^*$, as well as μ and $-\epsilon$.

These formulas have been given by Kottler[2] in the case of a time harmonic function. They were arrived at by means of the intermediary distribution of fictitious electric and "magnetic" current densities on the portion of S considered, and it is then necessary to ascertain that these expressions verify Maxwell's equations at point P. They are, in the present paper, directly derived from those equations and are obviously valid without any further verification.

The "Equivalence Theorem" is found useful in a number of problems. It shows that the experimental determination of the field distribution on any closed surface gives a means of computing the vector fields produced at any external point by sources that are located inside the surface. The same is true when theoretical considerations enable the determination of the fields on a surface S, or portion of said surface; this is generally an approximation only. This last has been found sufficient for the determination of radiation patterns of horns or reflectors used in conjunction with centimeter electromagnetic wavelengths.

Inversely, when the field distribution of secondary sources is assumed to be known on a closed surface S, said sources can be replaced by others that would give the same field distribution on S.

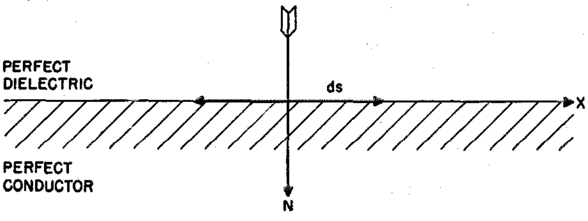

Fig. 2.

Consider, for instance, an unlimited plane supposedly a perfect conductor. Let an electromagnetic wave fall on an element ds of the plane in a medium supposed to be a perfect dielectric (Fig. 2). The magnetic inductive capacitance is the same in both media. Let $\boldsymbol{E}$, $\boldsymbol{H}$ be the incoming fields, and $\boldsymbol{E}'$, $\boldsymbol{H}'$ the fields due to the secondary sources induced in the reflecting plane. The boundary conditions give:

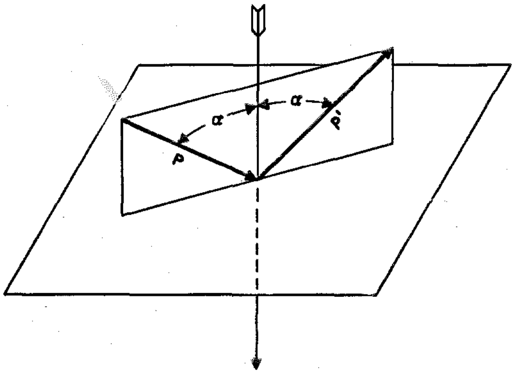

Fig. 3.

$$E'_x=-E_x,$$
$$E'_y=-E_y,$$
$$H'_z=-H_z.$$

There is no dissipation of energy: The flow of the incoming Poynting vector through ds should be equal to the flow of the Poynting vector associated with the reflected wave. It is easy to show that this leads to

$$H'_y=H_y,$$
$$H'_x=H_x.$$

Finally, both incoming and total fields, and consequently both incoming and reflected waves, verify Maxwell's equations. This can be shown to give

$$E'_z=E_z.$$

The association of the above conditions means that the total field $\mathcal{E}$, $\mathcal{H}$ on the reflector is such that

$$\mathcal{E}_z=2E_z,$$
$$\mathcal{E}_T=0,$$
$$\mathcal{H}_z=0,$$
$$\mathcal{H}_T=2H_T.$$

It also means that the incoming and reflected Poynting vectors are in the same plane and make an equal angle with the normal to the plane (Fig. 3).

The secondary sources can then be replaced by fictitious sources that would produce the fields E', H' at all points of the reflecting plane. This obviously leads to the well-known use of "electromagnetic images" to compute the total field in front of the unlimited reflector.

The expressions for the fields at point P given by the "Equivalence Theorem" are, of course, simpler when a time-harmonic function $e^{-i\omega t}$ is assumed. All vectorial operations with index P are then applied to an auxiliary function

$$\varphi = \frac{e^{-i\omega r/c}}{r}.$$

When the radiated fields are considered exclusively, only the terms in $1/r$ need be computed, and $1/r$ can be conventionally treated as a constant in the differentiations involved in the vectorial operations.

$TM_{0,1}$ Mode in Circular Wave Guides with Two Coaxial Dielectrics*

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FIELD components for a transverse magnetic wave in a wave guide with two coaxial dielectrics are computed. A typical example is given to show the calculation of guide dimensions to reduce phase velocity to a preassigned value.

• • •

A uniform circular wave guide may be conceived as a device that provides an interaction space between electrons and an electromagnetic field to transfer energy from one to the other. If electrons are injected axially in the guide and the field has an axial electric component, the important interaction is between the electrons and this axial component.

For unimpeded motion, the electrons must travel essentially in a vacuum and with a velocity necessarily less than that of light. On the other hand, the phase velocity of the electromagnetic field in an evacuated circular guide is always greater than the velocity of light. For effective interaction, this field must be slowed down, perhaps to the order of one-tenth or less of the velocity of light. This may be accomplished by surrounding the evacuated axial region of the guide with material of high dielectric constant. For simplicity, circular symmetry may be maintained.

An investigation of the propagation of an electromagnetic field with axial electric component in such a "loaded" guide has been made. Appropriate field equations, which take the necessary boundary conditions into account, are set up and solved. A relationship between the maximum value of axial electric component and transmitted power is given.

1. *Solution of Field Equations*

1.1 Differential Equations

Fig. 1 shows a cross section of a wave guide. The section of the axial region is a circle of radius a, the constants of the medium being μ and ϵ_2, the permeability and dielectric constants, respectively. The boundary of the guide is a perfect conductor of radius b. In the annular region between these two radii is a material of constants μ and ϵ_1. It is assumed that $\epsilon_1 > \epsilon_2$ always.

Cylindrical coordinates r, ϕ, and x will be used with the x axis coinciding with the axis of the guide and being positive in the direction toward a sink of power. For a transverse magnetic mode, take $H_x = 0$, and, in accordance with the conventional $TM_{0,1}$ mode for uniform dielectrics, specify additionally that $E_\phi = H_r = 0$. If a field satisfying these requirements exists in the guide under consideration, then the appropriate field equations in meter-kilogram-second units for either dielectric are of the following type.

$$\left.\begin{aligned} -\partial H_\phi/\partial x &= j\omega\epsilon E_r, \\ (1/r)(\partial/\partial r)(rH_\phi) &= j\omega\epsilon E_x, \\ (1/r)(\partial E_x/\partial\phi) &= 0, \\ (\partial E_r/\partial x) - (\partial E_x/\partial r) &= -j\omega\mu H_\phi, \\ (1/r)(\partial E_r/\partial\phi) &= 0. \end{aligned}\right\} \tag{1}$$

The subscript has been omitted from ϵ to avoid restricting it to either medium.

According to (1), E_r and E_x, therefore also H_ϕ, are independent of ϕ. The useful equations from (1) are

$$\left.\begin{aligned} \partial H_\phi/\partial x &= -j\omega\epsilon E_r, \\ (1/r)(\partial/\partial r)(rH_\phi) &= j\omega\epsilon E_x, \\ (\partial E_r/\partial x) - (\partial E_x/\partial r) &= -j\omega\mu H_\phi. \end{aligned}\right\} \tag{2}$$

Elimination of E_r and E_x among these equations leads to the following equation in H_ϕ:

$$\frac{\partial^2 H_\phi}{\partial x^2} + \frac{\partial}{\partial r}\left[\frac{1}{r}\frac{\partial}{\partial r}(rH_\phi)\right] = -\omega^2\mu\epsilon H_\phi = -\frac{\omega^2}{c^2}H_\phi, \tag{3}$$

where c is the velocity of propagation in a medium of unbounded extent with constants μ and ϵ.

Assume

$$H_\phi = R(r)e^{-\gamma x}, \tag{4}$$

where γ is the propagation constant in the x

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direction and R is a function of r to be determined. Substitution of (4) in (3) yields

$$\frac{d}{dr}\left[\frac{1}{r}\frac{d}{dr}(rR)\right]+\alpha^2R=0, \tag{5}$$

where

$$\alpha^2=\gamma^2+\omega^2/c^2. \tag{6}$$

Solutions of (5) are of the type

$$R=J_1(\alpha r),\ Y_1(\alpha r), \tag{7}$$

where J_1 and Y_1 are Bessel functions of the first and second kinds, respectively, and of the first order. If $Z_1(\alpha r)$ represents any linear combination of these functions, then

$$H_\phi=Z_1(\alpha r)e^{-\gamma x}. \tag{8}$$

For unattenuated transmission, γ is a pure imaginary; let $\gamma=j\beta$, with β real. Then

$$H_\phi=Z_1(\alpha r)e^{-j\beta x},\quad \alpha=[(\omega/c)^2-\beta^2]^{\frac{1}{2}}. \tag{9}$$

From (2), we get immediately

$$\left.\begin{aligned} E_r&=(\beta/\omega\epsilon)Z_1(\alpha r)e^{-j\beta x},\\ E_x&=(\alpha/j\omega\epsilon)Z_0(\alpha r)e^{-j\beta x},\end{aligned}\right\} \tag{10}$$

where Z_0 is a linear combination of Bessel functions of the first and second kind, zeroth order, and we have made use of the relation[1]

$$\frac{d}{dr}[rZ_1(\alpha r)]=\alpha rZ_0(\alpha r). \tag{11}$$

J_0 and J_1 are finite and continuous everywhere in the complex plane, but Y_0 and Y_1 have infinite discontinuities at $r=0$. These latter functions are therefore inadmissable in the axial region of Fig. 1. The solutions therefore run as follows.

In medium 1,

$$\left.\begin{aligned} H_{\phi1}&=[A_1J_1(\alpha_1r)+B_1Y_1(\alpha_1r)]e^{-j\beta_1x},\\ E_{r1}&=\frac{\beta_1}{\omega\epsilon_1}[A_1J_1(\alpha_1r)+B_1Y_1(\alpha_1r)]e^{-j\beta_1x},\\ E_{x1}&=\frac{\alpha_1}{j\omega\epsilon_1}[A_1J_0(\alpha_1r)+B_1Y_0(\alpha_1r)]e^{-j\beta_1x}.\end{aligned}\right\}$$

$$a\leqq r\leqq b. \tag{12a}$$

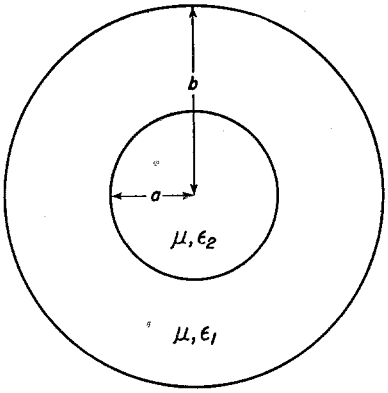

Fig. 1—Cross section of wave guide having two coaxial dielectrics.

In medium 2,

$$\left.\begin{aligned} H_{\phi2}&=A_2J_1(\alpha_2r)e^{-j\beta_2x},\\ E_{r2}&=(\beta_2/\omega\epsilon_2)A_2J_1(\alpha_2r)e^{-j\beta_2x},\\ E_{x2}&=(\alpha_2/j\omega\epsilon_2)A_2J_0(\alpha_2r)e^{-j\beta_2x},\end{aligned}\right\}0\leqq r\leqq a. \tag{12b}$$

A_1, B_1, and A_2 are constants to be determined from the boundary and power conditions, α_1 and α_2 are the propagation constants, and β_1 and β_2 are the phase constants, in media 1 and 2, respectively.

1.2 Boundary Conditions

Matching of tangential components of the electric field at the dielectric interface requires that

$$E_{x1}=E_{x2}\quad\text{at}\quad r=a$$

for all values of x; i.e.,

$$\frac{\alpha_1}{j\omega\epsilon_1}[A_1J_0(\alpha_1a)+B_1Y_0(\alpha_1a)]e^{-j\beta_1x}=\frac{\alpha_2}{j\omega\epsilon_2}A_2J_0(\alpha_2a)e^{-j\beta_2x}. \tag{13}$$

This can be satisfied for all values of x if, and only if, $\beta_1=\beta_2=\beta$. The phase constant (hence the velocity of propagation) is the same in both media, as might have been expected.

$$\alpha_1^2-\omega^2/c_1^2=-\beta^2=\alpha_2^2-\omega^2/c_2^2. \tag{14}$$

Assume the phase velocity V_p to be intermediate between c_1 and c_2

$$c_1<V_p<c_2,\quad V_p=\omega/\beta. \tag{15}$$

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[1] Jahnke and Emde, "Tables of Functions," 3rd edition, B. G. Teubner, Leipzig, 1938; p. 145.

This leads to the inequality

$$\alpha_2^2 < 0 < \alpha_1^2, \qquad (16)$$

whence α_1 is real while α_2 is a pure imaginary. The field in medium 2 therefore behaves according to Bessel functions of the first kind with pure imaginary argument. Curves for $J_0(jz)$ and $-jJ_1(jz)$, frequently designated as $I_0(z)$ and $I_1(z)$, respectively, have been published.[2] The function $I_0(z)$ is particularly interesting since it shows the radial variation of the axial electric field. This field is fairly uniform for sufficiently small z. For $z<0.6$, the variation is less than 10 percent; for $z<2$, the variation is less than about 2:1.

Next, for the tangential component of electric field to vanish at the conductor surface, it is required that $E_{x1}=0$ at $r=b$, i.e.,

$$A_1 J_0(\alpha_1 b) + B_1 Y_0(\alpha_1 b) = 0. \qquad (17)$$

Matching the normal components of dielectric flux at the dielectric interface requires that

$$\epsilon_1 E_{r1} = \epsilon_2 E_{r2} \quad \text{at} \quad r=a; \quad \text{i.e.,}$$

$$A_1 J_1(\alpha_1 a) + B_1 Y_1(\alpha_1 a) = A_2 J_1(\alpha_2 a). \qquad (18)$$

Matching the tangential magnetic field at the dielectric interface yields no further information. Collecting and rearranging (13), (17), and (18), we have the set of linear homogeneous equations in A_1, B_1, and A_2:

$$\left.\begin{aligned} \frac{\alpha_1}{\epsilon_1} J_0(\alpha_1 a) A_1 + \frac{\alpha_1}{\epsilon_1} Y_0(\alpha_1 a) B_1 \qquad & \\ - \frac{\alpha_2}{\epsilon_2} J_0(\alpha_2 a) A_2 &= 0, \\ J_0(\alpha_1 b) A_1 + Y_0(\alpha_1 b) B_1 &= 0, \\ J_1(\alpha_1 a) A_1 + Y_1(\alpha_1 a) B_1 - J_1(\alpha_2 a) A_2 &= 0 \end{aligned}\right\} (19)$$

A necessary and sufficient condition that these equations be consistent is that the determinant of the coefficients vanish:

$$\begin{vmatrix} \frac{\alpha_1}{\epsilon_1} J_0(\alpha_1 a) & \frac{\alpha_1}{\epsilon_1} Y_0(\alpha_1 a) & -\frac{\alpha_2}{\epsilon_2} J_0(\alpha_2 a) \\ J_0(\alpha_1 b) & Y_0(\alpha_1 b) & 0 \\ J_1(\alpha_1 a) & Y_1(\alpha_1 a) & -J_1(\alpha_2 a) \end{vmatrix} = 0. \qquad (20)$$

If β is specified in advance, then α_1 and α_2 are known from (14), and (20) specifies a necessary and sufficient relationship between a and b. The manner of using this information will be illustrated by a later example.

With (20) satisfied, any two of the three relations given in (19) may be used to solve for the ratios of the constants A_1, B_1, and A_2. By using the first and the third,

$$A_1 = K_A A_2, \quad B_1 = K_B A_2, \qquad (21)$$

where

$$\left.\begin{aligned} K_A &= -\frac{\pi \epsilon_1 a}{2} \begin{vmatrix} \frac{\alpha_2}{\epsilon_2} J_0(\alpha_2 a), & \frac{\alpha_1}{\epsilon_1} Y_0(\alpha_1 a) \\ J_1(\alpha_2 a), & Y_1(\alpha_1 a) \end{vmatrix} \\ K_B &= -\frac{\pi \epsilon_1 a}{2} \begin{vmatrix} \frac{\alpha_1}{\epsilon_1} J_0(\alpha_1 a), & \frac{\alpha_2}{\epsilon_2} J_0(\alpha_2 a) \\ J_1(\alpha_1 a), & J_1(\alpha_2 a) \end{vmatrix} \end{aligned}\right\} (22)$$

and we have made use of the standard identity

$$Y_0(s) J_1(s) - Y_1(s) J_0(s) = 2/\pi s.$$

The field components are then given as follows:

Medium 1: $a \leqq r \leqq b$.

$$\left.\begin{aligned} E_{x1} &= A_2 \frac{\alpha_1}{j\omega\epsilon_1} [K_A J_0(\alpha_1 r) + K_B Y_0(\alpha_1 r)] e^{-j\beta x}, \\ E_{r1} &= A_2 \frac{\beta}{\omega\epsilon_1} [K_A J_1(\alpha_1 r) + K_B Y_1(\alpha_1 r)] e^{-j\beta x}, \\ H_{\phi 1} &= A_2 [K_A J_1(\alpha_1 r) + K_B Y_1(\alpha_1 r)] e^{-j\beta x}, \end{aligned}\right\} (23a)$$

Medium 2: $0 \leqq r \leqq a$.

$$\left.\begin{aligned} E_{x2} &= A_2 \frac{\alpha_2}{j\omega\epsilon_2} J_0(\alpha_2 r) e^{-j\beta x}, \\ E_{r2} &= A_2 \frac{\beta}{\omega\epsilon_2} J_1(\alpha_2 r) e^{-j\beta x}, \\ H_{\phi 2} &= A_2 J_1(\alpha_2 r) e^{-j\beta x}. \end{aligned}\right\} (23b)$$

In (23), the constant A_2 remains to be determined.

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[2] Page 224 of reference 1.

1.3 Axial Field as a Function of Transmitted Power

The constant A_2 may be evaluated in terms of the peak transmitted power P_M or twice the average power P. For this purpose, the complex Poynting vector is integrated over a convenient cross section of the guide. This reduces to

$$P_M = \int_0^b \int_0^{2\pi} E_r H_\phi^* \cdot r d\phi dr,$$

where H_ϕ^* is the complex conjugate of H_ϕ. By (22), K_A and K_B both contain the factor α_2 and are therefore pure imaginaries. If E_r and H_ϕ are specified to have zero phase angle at $x=0$, then A_2 is also a pure imaginary, and j appears only in the exponential factor. The integration then runs

$$P_M = 2\pi A_2^2 \frac{\beta}{\omega\epsilon_1} F(a,b;\epsilon_1,\epsilon_2),$$

where

$$F = \frac{\epsilon_1}{\epsilon_2} \int_0^a r J_1^2(\alpha_2 r) dr + \int_a^b r[K_A J_1(\alpha_1 r) + K_B Y_1(\alpha_1 r)]^2 dr \tag{24}$$

or

$$A_2 = ((P_M/2\pi)(\omega\epsilon_1/\beta F))^{\frac{1}{2}}. \tag{25}$$

At the center of the tube, the maximum value of the axial field E_{x2} is

$$E_{x0} = \left| A_2 \frac{\alpha_2}{\omega\epsilon_2} \right| = \left| \frac{\alpha_2}{\omega\epsilon_2} \left(\frac{P_M}{2\pi} \frac{\omega\epsilon_1}{\beta F} \right)^{\frac{1}{2}} \right|. \tag{26}$$

1.4 Further Discussion of the Solution

The determinant (20) may be considered to yield a solution for b when α_1 and α_2 are known. After defining a set of functions

$$Q_1(s) = \frac{J_0(s)}{Y_0(s)}, \quad Q_2(s) = -j\frac{J_1(js)}{J_0(js)}, \quad Q_3(s) = \frac{J_0(s)}{J_1(s)}, \quad Q_4(s) = \frac{Y_1(s)}{J_1(s)}, \tag{27}$$

we can reduce (20) to the form

$$Q_1(\alpha_1 b) = \frac{1 - j\dfrac{\alpha_1}{\alpha_2}\dfrac{\epsilon_2}{\epsilon_1} Q_2(\alpha_2 a) Q_3(\alpha_1 a)}{Q_4(\alpha_1 a) - j\dfrac{\alpha_1}{\alpha_2}\dfrac{\epsilon_2}{\epsilon_1} Q_2(\alpha_2 a) \dfrac{Q_3(\alpha_1 a)}{Q_1(\alpha_1 a)}}. \tag{28}$$

All of these Q functions, with the exception of $Q_2(s)$, have been plotted.[3] $Q_2(s)$ is plotted in Fig. 2 of this paper. The use of these curves simplifies the calculation of b.

The chief remaining problem is an evaluation of the function $F(a, b; \epsilon_1, \epsilon_2)$ in (24). The function is integrable in known Bessel functions. By reference to a standard treatise on Bessel

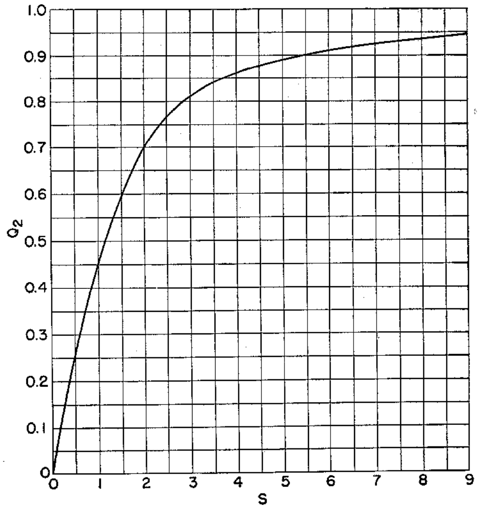

Fig. 2—A useful ratio of certain Bessel functions of imaginary arguments. $Q_2 = -j[J_1(js)/J_0(js)]$.

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[3] Pages 200–203 of reference 1.

functions,[4] it can be shown that

$$F=\frac{1}{\alpha_2^2}F_1(\alpha_2a)+\frac{K_A^2}{\alpha_1^2}[F_1(\alpha_1b)-F_1(\alpha_1a)]+\frac{2K_AK_B}{\alpha_1^2}[F_3(\alpha_1b)-F_3(\alpha_1a)]+\frac{K_B^2}{\alpha_1^2}[F_2(\alpha_1b)-F_2(\alpha_1a)], \quad (29)$$

where

$$\left.\begin{aligned} F_1(s)&=\frac{s^2}{2}\{[J_1(s)]^2-J_0(s)J_2(s)\},\\ F_2(s)-F_2(u)&=\frac{s^2}{2}\{[Y_1(s)]^2-Y_0(s)Y_2(s)\}-\frac{u^2}{2}\{[Y_1(u)]^2-Y_0(u)Y_2(u)\},\\ F_3(s)&=\tfrac{1}{2}\{s^2[Y_1(s)J_1(s)+Y_2(s)J_2(s)]-s[Y_1(s)J_2(s)+Y_2(s)J_1(s)]\}.\end{aligned}\right\} \quad (30)$$

Future work may indicate whether or not these results can be simplified.

2. *Typical Example*

Suppose a $TM_{0,1}$ wave is to be set up at $\omega=3\cdot10^{10}$ per second traveling at one-tenth the velocity of light in vacuum, so that $V_p=3\cdot10^7$ meters per second, whence $\beta=(10)^3$ per meter. Assume the dielectric material is such that

$\mu=4\pi10^{-7}$ henry per meter,
$\epsilon_1=(10^{-7}/18\pi)$ farad per meter
(relative dielectric constant $=200$),
$\epsilon_2=(10^{-9}/36\pi)$ farad per meter.

Then $\alpha_1=10^3$, $\alpha_2=j10^3$ by (14). Next, let us try using a radius $a=2$ millimeters $=2\cdot10^{-3}$ meter. Then

$$\begin{aligned}\alpha_1a&=2,\\ \alpha_2a&=j2,\\ \alpha_1/\epsilon_1&=5.55\cdot10^{11},\\ \alpha_2/\epsilon_2&=j1.11\cdot10^{14}.\end{aligned}$$

From Fig. 2, $Q_2(j2)=0.70$, and from reference 1 and (27),

$$\begin{aligned}Q_1(2)&=0.45,\\ Q_3(2)&=0.39,\\ Q_4(2)&=-0.19.\end{aligned}$$

Then by (28), $Q_1(\alpha_1b)=-5.2$ whence

$$\alpha_1b=0.73,\ 3.78,\ 6.89,\ \ldots.$$

and

$$b=0.7,\ 3.8,\ 6.9,\ \ldots,\ \text{millimeters}.$$

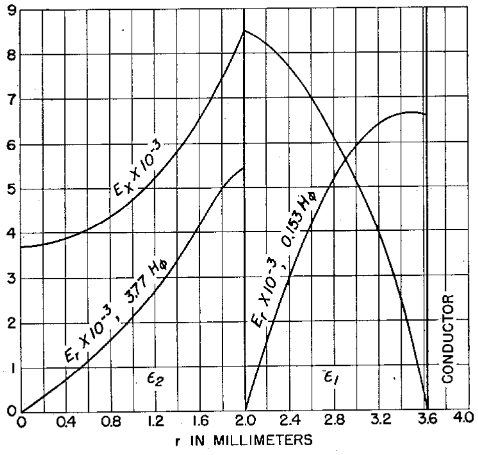

Fig. 3—Distribution of field components in a typical example.

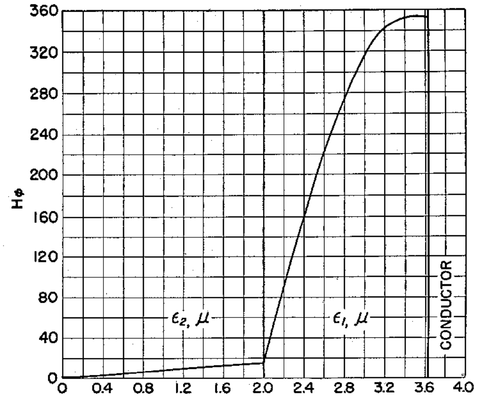

Fig. 4—Distribution of magnetic intensity in a typical example.

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[4] G. N. Watson, "A Treatise on the Theory of Bessel Functions," 2nd edition, Macmillan Company, New York; 1944.

We take $b=3.8$ millimeters as the smallest available value of $b>a$. A more careful calculation using tables rather than curves shows that $b=3.6$ millimeters.

Next, compute K_A and K_B using (22).

$$K_A=j153.3,$$
$$K_B=j812.5.$$

The expressions for the field components (23) then become

$$E_{x1}=A_2 2842[J_0(\alpha_1 r)+5.30\,Y_0(\alpha_1 r)]e^{-i\beta x},$$
$$E_{r1}=jA_2 2892[J_1(\alpha_1 r)+5.30\,Y_1(\alpha_1 r)]e^{-i\beta x},$$
$$H_{\phi 1}=jA_2 153.3[J_1(\alpha_1 r)+5.30\,Y_1(\alpha_1 r)]e^{-i\beta x},$$
$$E_{x2}=A_2 3703 J_0(\alpha_2 r)e^{-i\beta x},$$
$$E_{r2}=A_2 3774 J_1(\alpha_2 r)e^{-i\beta x},$$
$$H_{\phi 2}=A_2 J_1(\alpha_2 r)e^{-i\beta x}.$$

These quantities are plotted as functions of r in Figs. 3 and 4. Here it is plain that the electric vectors have the same orders of magnitude in both dielectrics, while the magnetic vector is practically negligible in the central dielectric. This fact, combined with the small cross-sectional area of the inner dielectric, results in most of the transmitted power being propagated in the outer dielectric.

The factor F in (24) may be evaluated by numerical integration, whence it turns out that $F=-0.00037-0.3475=-0.3479$, the smaller term being the contribution of the inner dielectric. Since by (26) the transmitted power is related linearly to F, it can be seen that only about 0.1 percent of the power is propagated in the central region.

By substitution in (26) we now have immediately $E_{x0}=5.77(P_M)^{\frac{1}{2}}$, volts per centimeter, at the axis of the guide.

Discussion

Remarks on Slow Waves in Cylindrical Guides*

SEVERAL articles, one in 1944 [1] and two more recently, [2] have appeared on the question of concentric dielectrics in circular wave guides, each treating it from a different point of view. The problem considered by the more recent articles is the utilization of the $TM_{0,1}$ mode in a circular wave guide with two coaxial dielectrics to obtain wave propagation with a phase velocity below that of light velocity. A main objective of each of these articles is to obtain a method whereby one is able to determine the necessary values of the dielectric constants and radii involved, when the phase velocity is given some pre-assigned value. The earlier article considers the problem more generally. All of the authors approach the problem by solving for the fields and then applying the appropriate boundary conditions.

The purpose of this letter is twofold; (1) to indicate that an alternative and simpler method may be used to derive the necessary relations, and (2) to acquaint the reader with the existence of certain curves whose use greatly eases the calculations.

The properties of an ideal wave guide are determined once the cut-off frequency is determined. The cut-off frequency for a given mode corresponds to resonance for waves propagating transversely in the cross section. Therefore, to find the cut-off frequency of the guide, it is necessary only to determine the resonant frequency of the two-dimensional problem defined by the guide cross section. The condition for resonance that we apply is that, at any radius, the admittances looking in opposite radial directions should be equal and opposite.

Since the above-mentioned articles do not agree as to notation, it is in place to specify the notation used here. The inner and outer regions shall be regions 1 and 2, respectively (see Fig. 1). Dielectric constant ϵ_2 is greater than ϵ_1. k_c shall represent the cut-off wave number, and β, the propagation constant in the axial direction. We then have:

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* By A. A. Oliner, Microwave Research Institute, Polytechnic Institute of Brooklyn, Brooklyn, New York, reprinted from *Journal of Applied Physics*, v. 19, pp. 109–110; January, 1948.

[1] L. Pincherle, "Electromagnetic Waves in Metal Tubes Filled Longitudinally with Two Dielectrics," *Physical Review*, v. 66, p. 118; 1944.

[2] Sidney Frankel, "$TM_{0,1}$ Mode in Circular Wave Guides with Two Coaxial Dielectrics," *Journal of Applied Physics*, v. 18, p. 650; 1947; G. G. Bruck and E. R. Wicher, "Slow Transverse Magnetic Waves in Cylindrical Guides," *Journal of Applied Physics*, v. 18, p. 766; 1947.

$$k_{c1}=(k^2\epsilon_1-\beta^2)^{\frac{1}{2}}, \quad k_{c2}=(k^2\epsilon_2-\beta^2)^{\frac{1}{2}}, \tag{1}$$

where $k=\omega/c$ and $\beta=\omega/v_p$, v_p being the phase velocity.

A radial guide is the appropriate one to use in the cross section since we have propagation down a circularly cylindrical wave guide. The expression for the admittance $Y(r_1)$ at radius r_1 in terms of the admittance $Y(r_2)$ at r_2, for the lowest E-type mode in a radial guide can be written in analogy with the uniform line case as:[3]

$$Y(r_1)=Y_0(r_1)\frac{j+\zeta(x,\,y)ct(x,\,y)\,Y(r_2)/Y_0(r_2)}{Ct(x,\,y)+j\zeta(x,\,y)\,Y(r_2)/Y_0(r_2)}, \tag{2}$$

where

$$\begin{aligned} \zeta(x,\,y)&=\frac{J_0(x)N_0(y)-N_0(x)J_0(y)}{J_1(x)N_1(y)-N_1(x)J_1(y)},\\ ct(x,\,y)&=\frac{J_1(x)N_0(y)-N_1(x)J_0(y)}{J_0(x)N_0(y)-N_0(x)J_0(y)},\\ Ct(x,\,y)&=\frac{J_1(y)N_0(x)-N_1(y)J_0(x)}{J_1(x)N_1(y)-N_1(x)J_1(y)}, \end{aligned} \tag{3}$$

$x=k_cr_1$, $y=k_cr_2$, and Y_0 is the characteristic admittance of the guide which is a function of the radius. It may be seen that if a short circuit exists at r_2, then $Y(r_2)=\infty$, and $Y(r_1)$ reduces to:

$$\overrightarrow{Y}(r_1)=-jY_0(r_1)ct(x,\,y). \tag{4}$$

Similarly, we find that for any finite $Y(r_2=0)$:

$$\overleftarrow{Y}(r_1)=jY_0(r_1)\frac{J_1(x)}{J_0(x)}. \tag{5}$$

The condition for resonance is satisfied if, at radius $r=r_1$, we set $\overrightarrow{Y}+\overleftarrow{Y}=0$. We then obtain:

$$Y_{01}(r_1)\frac{J_1(k_{c1}r_1)}{J_0(k_{c1}r_1)}=Y_{02}(r_1)ct(x,\,y), \tag{6}$$

where $x=k_{c2}r_1$ and $y=k_{c2}r_2$. For E-type modes the ratio of the characteristic admittances is given by

$$\frac{Y_{01}(r_1)}{Y_{02}(r_1)}=\frac{\epsilon_1k_{c2}}{\epsilon_2k_{c1}}. \tag{7}$$

We may note that k_c is the propagation constant in the transverse region, although β is the propagation constant in the axial direction. Substituting for the characteristic admit-

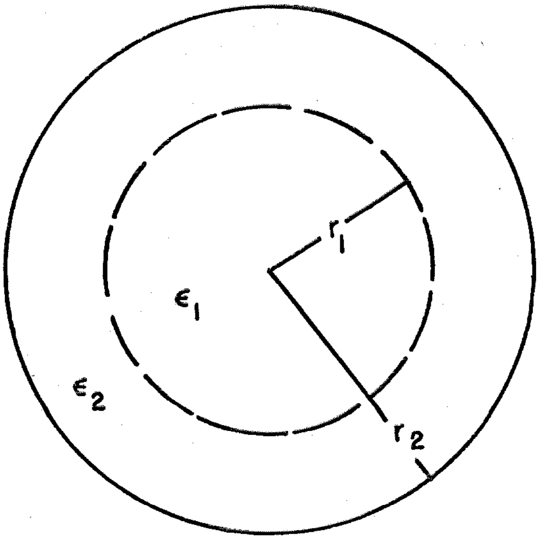

Fig. 1.

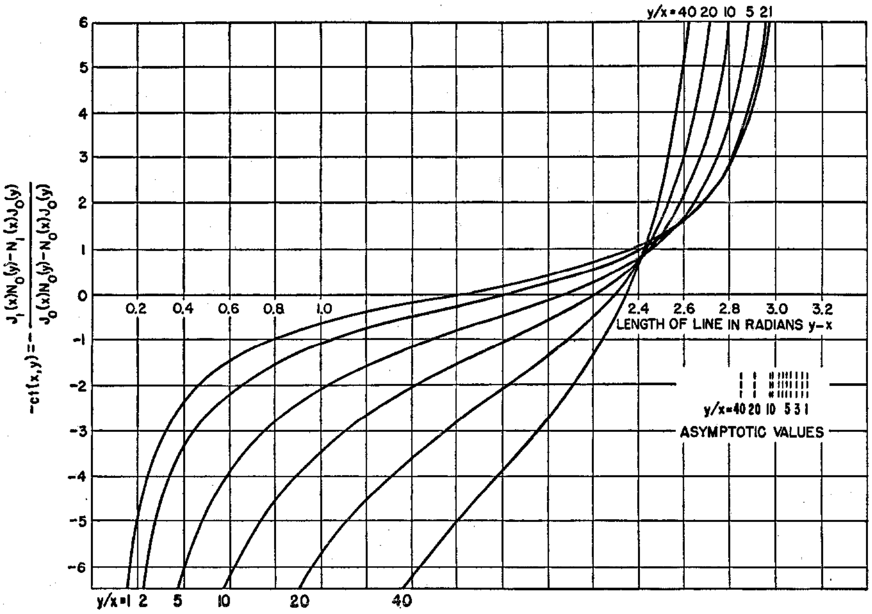

Fig. 2.

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[3] Montgomery, Dicke, and Purcell, "Principles of Microwave Circuits" (v. 8, Radiation Laboratory Series), McGraw-Hill Book Company, New York, New York, to be published; a brief derivation may be found in Chapter 8.

tances, we obtain finally:

$$\frac{\epsilon_1 k_{c2}}{\epsilon_2 k_{c1}} \frac{J_1(k_{c1}r_1)}{J_0(k_{c1}r_1)} = ct(x, y), \qquad (8)$$

where $x = k_{c2}r_1$ and $y = k_{c2}r_2$.

It may be seen that (8) is identical, except for notation, with (8) of the article by Bruck and Wicher. These authors state that this expression is the relation between the field velocity, tube radii, and dielectric constant, from which they obtain their results. They further state that this relation "can be solved by a combination of patience with a variety of numerical procedures." However, there exist curves of the combination of Bessel functions represented by $ct(x, y)$. (See Fig. 2.) The values of $ct(x, y)$ are plotted against $y-x$, with y/x as a parameter. The results desired may be obtained quite rapidly by the use of these *ct* curves. The method involved here is essentially the same as that used by these authors; the curves simply serve to facilitate the calculations.

As an example, one of the points considered by the above authors was checked using the *ct* curves. The phase velocity chosen was $v_p = 0.2c$, and the smaller and larger radii were 0.05λ and 0.10λ, respectively. The required ratio of dielectric constants must be determined. The value $\epsilon_2/\epsilon_1 = 57.3$ was obtained by means of the *ct* curves. Bruck and Wicher state that $\epsilon_2/\epsilon_1 = 57.25$.

In the article by Frankel, the final relations are obtained in somewhat different form. However, after appropriate rearranging of terms, (28) of Frankel's article may be seen to be identical, again except for notation, with (8) above. In his "typical example," Frankel specifies the values of all the necessary parameters except the larger radius, which must be solved for. His result, obtained graphically, is 3.8 millimeters. The result obtained by means of the *ct* curves is 3.77 millimeters. Frankel states further that "a more careful calculation using tables rather than curves" shows that the result should be 3.6 millimeters. The writer believes that this statement is in error as the use of tables with (8) above yields a value of 3.773 millimeters.

High-Ratio Multivibrator Frequency Divider

By M. SILVER and A. SHADOWITZ

Federal Telecommunication Laboratories, Incorporated, Nutley, New Jersey

FREQUENCY-DIVIDER chains of considerable length for obtaining two widely separated frequencies in synchronism with each other have been increasingly used in recent years. In television practice, for example, it is necessary to generate a 60-cycle-per-second signal in exact lock with another of 31,500 cycles.[1] In frequency-modulation broadcast transmitters of the direct type, it is usual to divide the master-oscillator frequency by a factor of several hundred in order to reduce the index of modulation before comparison with the crystal oscillator. It has been customary to use a comparatively large number of dividers, either of the multivibrator or blocking-oscillator type, to accomplish the above, since each divider must be restricted to a rather small order of division, about 5, for stability under varying operating conditions. A system[2] has also been designed for accomplishing high-order division with only a few stages, by using saw-tooth dividers. The present article describes a somewhat similar circuit for a high-ratio multivibrator divider.

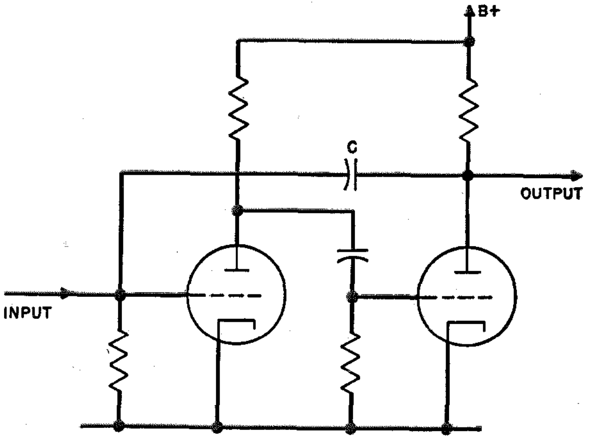

Fig. 1—Typical multivibrator-divider circuit.

1. Circuit Operation

1.1 Simple Lock-In Multivibrator

Fig. 1 is the circuit diagram of an ordinary multivibrator divider. A synchronizing voltage (either pulse or sine wave) applied at the input will lock the multivibrator output frequency at an exact submultiple of the synchronizing frequency if the time constants and the amplitude of the input voltage are properly chosen. It is difficult to stabilize the order of division with this circuit; instead of dividing, say, by 20/1 as desired, a slight change in any one of a number of factors may cause division by 19/1 or 21/1. The value of the plate-supply voltage, the synchronizing-voltage amplitude or frequency, and aging of the tubes all combine to make high-order division critical.

1.2 Insertion of Clippers

Suppose that we remove capacitor C (Fig. 1) between the first grid and second plate, and replace it with a tuned circuit followed by a two-stage clipper, as shown in Fig. 2. The output of the second clipper is fed back to the first grid. Resistor R has a high value in order that the Q of the tuned circuit may be high, and the bias on the first clipper stage is made high enough to prevent the flow of grid current. The capacitors connected with the clippers are all comparatively large.

It can be seen that this circuit will oscillate only in the immediate vicinity of the resonant frequency of the tuned circuit, with or without synchronizing voltage at the input. A sine wave will be obtained at output A, while one side of the sine wave will be clipped at the grid of the second clipper stage. At output B (and in the multivibrator circuit proper), a square wave will be obtained.

If we introduce a synchronizing voltage of a frequency close to an integral multiple of the tuned-circuit frequency, the multivibrator will

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[1] A. Rufus Applegarth, "Synchronizing Generators for Electronic Television," *Proceedings of the I.R.E.*, v. 34, pp. 128W–139W; March, 1946.

[2] Geoffrey Builder, "A Stabilized Frequency Divider," *Proceedings of the I.R.E.*, v. 29, pp. 177–181; April, 1941.

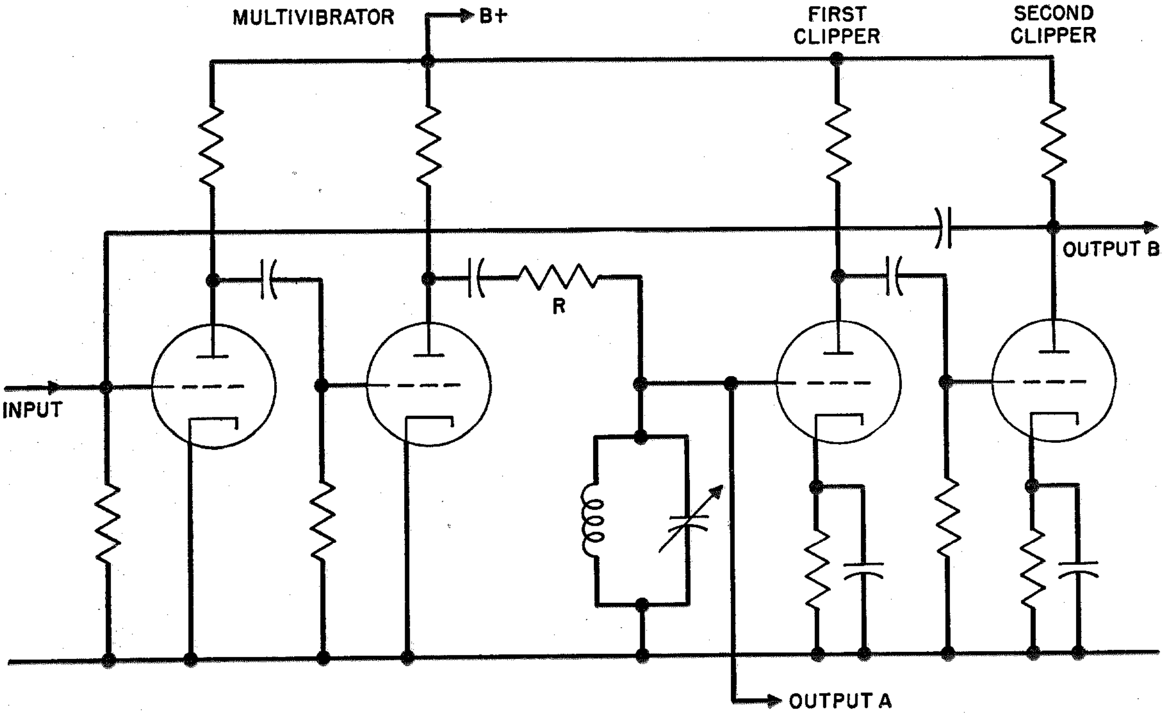

Fig. 2—Modified circuit of Fig. 1, in which capacitor *C* is replaced by a tuned circuit and two clipper stages. Larger and more stable division ratios between input and output frequency are possible with this circuit.

adjust itself to an exact subharmonic of the input.

The action here is that of a lock-in of an unstable oscillator to a stable oscillator; a necessary condition for lock-in is that the two oscillators should have a harmonic frequency in common, and this is the reason for the double-clipper stage, which generates harmonics by squaring the sine wave. The frequency of the synchronizing voltage (the stable oscillator) must be close to one of the multivibrator harmonics for lock-in to occur; the higher the order of division desired, the more important it is that the waves be square. Since it is difficult to obtain good square waves at high repetition rates, it is also more difficult to obtain good division at the higher frequencies. The actual lock-in forces are produced by third-order nonlinearities in the tube operating characteristic.

1.3 Lock-In Range

For any given order of division, there is a range of synchronizing frequencies at which proper division is obtained. An example is given in Fig. 3. This synchronizing-frequency range is directly proportional to the multivibrator operating range, which is determined by the *Q* of the tuned

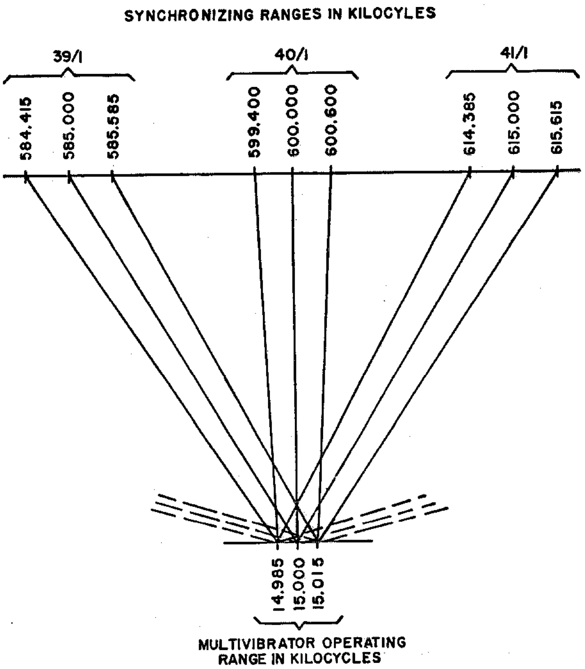

Fig. 3—Illustration of three of the synchronizing ranges of the multivibrator-divider circuit of Fig. 2.

circuit. If the multivibrator is driven at a different frequency from the exact resonant frequency of the tuned circuit, a phase shift is introduced in the tuned circuit; the phase shift for a given deviation in frequency increases with increasing Q, and proper division no longer occurs when the phase shift exceeds a certain critical value. Thus for higher orders of division, where adjacent synchronizing ranges become closer to each other, it is necessary to restrict the multivibrator operating range to smaller regions by using higher Q's.

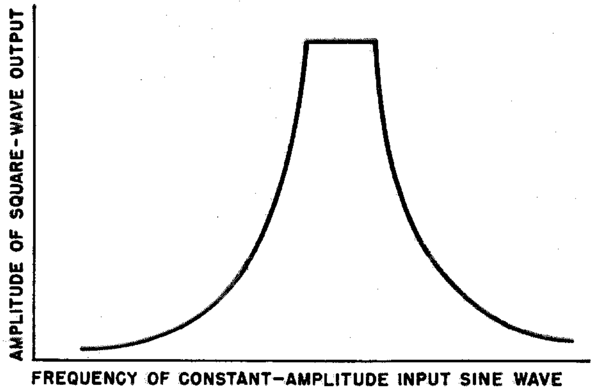

Fig. 4—Transmission characteristic of tuned-circuit-and-clipper coupling circuit. It exhibits the typical frequency-versus-amplitude curve of a tuned circuit, with a plateau at the top due to the clippers.

The transmission characteristic of the circuit is shown in Fig. 4. This curve may be obtained by removing the capacitor between the input grid and the second clipper plate, applying a variable-frequency constant-amplitude sine wave to the input, and examining the output at the second clipper plate. The output voltage will reach the maximum value at very low input voltages (of the order of millivolts), and will then be a square wave. The plateau width (Fig. 4) can be varied by changing the bias of the first clipper, and bears no relation to the multivibrator lock-in range.

2. *Final Circuit*

Fig. 5 is the detailed schematic of a circuit that gives stable division ratios as high as 300/1 (4.5 megacycles to 15 kilocycles). Adjustment for proper division can be made most easily with an oscilloscope at any of the plates; when the multivibrator locks with the synchronizing frequency, it is possible to count the number of synchronizing-frequency pulses per multivibrator cycle.

3. *Conclusion*

The circuit described has the one outstanding advantage of simplicity; it requires but one tuned circuit and two tubes, and is quite insensitive to changes in operating conditions. In fact, since the values of resistance and capacitance have so little effect on the frequency, it would seem incorrect to call the circuit a multivibrator, except that it is derived from the ordinary multivibrator, and because it produces easily synchronized square waves.

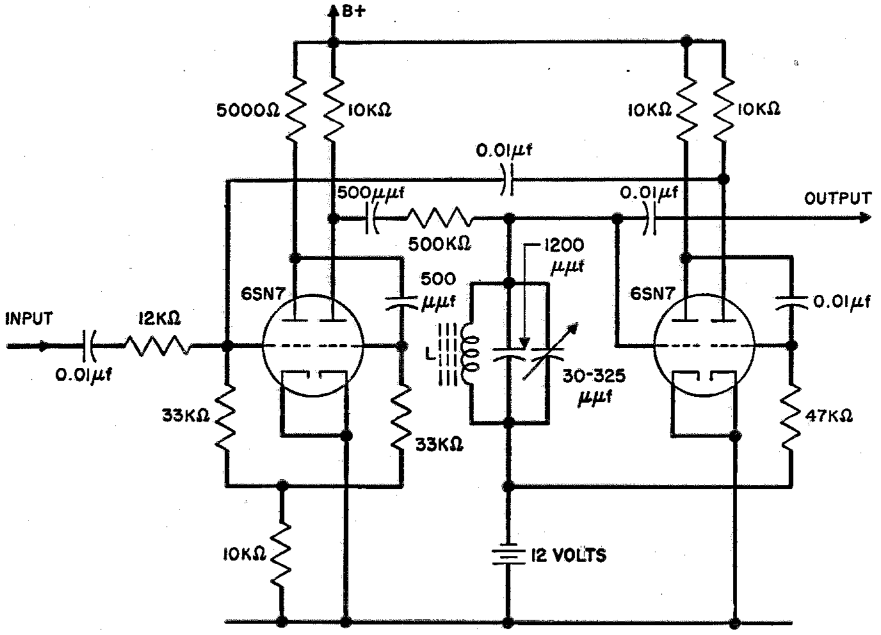

Fig. 5—Circuit dividing from 4.5 megacycles to 15 kilocycles, a 300-to-1 range. Inductance *L* consists of 800 turns of 28-American-Wire-Gauge wire on a powdered-iron toroid.

Thermo-Electric and Conductive Properties of Blue Titanium Dioxide*

By H. K. HENISCH

Department of Physics, University of Reading, England

1. *Introduction*

THE thermo-electric properties of blue semi-conducting titanium dioxide are of special interest, both in connection with the possible practical applications of this material and because of the theoretical importance of these effects. Titanium dioxide is also in many ways an ideal semi-conductor because its conductivity is purely electronic up to quite high temperatures[1] and the material is entirely stable under all normal conditions. Moreover, as far as is known, its structure with regard to electronic-energy distribution is substantially in accordance with the simple concepts on which semi-conductor theories are usually based. Experiments on this material should, therefore, offer welcome opportunities for checking the validity of these theories.

Most theoretical work on semi-conductors necessarily assumes that the properties concerned are those of a single crystal of the material. During practical experiments, this condition is rarely satisfied. Some experiments have indeed been carried out on single crystals and have yielded results of great value,[2,3] but most semi-conductors otherwise available for investigation are in the form of micro-crystalline conglomerates. This means that properties like the conductivity will depend not only on the characteristics of individual grains, but also on the shape, orientation, and packing of grains within the conglomerate as a whole. As a rule, it is not possible to control these two contributing factors independently during the preparation of specimens, nor is it possible to distinguish between them by conductivity measurements alone. To test the predictions of theory, it is, therefore, necessary to measure some property that depends only on the composition (impurity content) and not on the grain structure. The thermo-electric and Hall effects are two such properties and of these the former is likely to be the simpler to measure. Thermo-electric measurements are thus of value as means not only of testing the theory of semi-conductors, but also (in combination with conductivity measurements on the same sample) of enabling the relative importance of composition and grain structure to be assessed.

A considerable amount of experimental and theoretical work has been carried out on this subject in the past, probably most of it with special reference to cuprous oxide,[4–13] but some also in connection with selenium,[14] lead sulphide,[15] stannous sulphide,[16] germanium,[17,18] cadmium oxide,[19] silicon,[20] zinc oxide,[21] and many other substances and mixtures.[22] Expressions for the thermo-electric power and the total thermo-electromotive force have been developed in terms of the temperatures T_1 and T_2 of two metal–semi-conductor junctions and various parameters characteristic of the materials concerned.[10,17,23] In practice, many of these equations cannot readily be subjected to experimental test, because they are complicated and contain terms that cannot be independently estimated. It will be shown that by suitable choice of parameters it is possible to deduce approximate equations that are more suitable for direct experimental check, and the application of which leads to results of practical interest.

As is well known, the temperature dependence of the conductivity is given by an equation

$$\sigma = \sigma_0 \exp\left[-eV_a/2kT\right],$$

where σ_0 usually varies much more slowly with temperature than the exponential term. According to the simplest theory, eV_a is the energy required to free an electron from an impurity centre and raise it into the conduction band. This activation energy also enters into the thermo-electric equations. It is shown in the present paper that the experimental values for this energy derived from conductivity and thermo-electric measurements on sintered titanium dioxide specimens agree only as regards order of magnitude. This would be expected for specimens that are not entirely homogeneous, in the sense that their impurity content and activation energy are subject to local variations. It is suggested that while the impurity content can vary continuously, only a few discrete values of the activation energy are possible and the different grains each have one or other of these values. The measured activation energy thus represents some form of average. The effect of non-uniform impurity content on the thermoelectric properties of the specimens can be calculated for an ideal case in which the impurity concentration varies continuously throughout the solid. The conclusions are supported by the experimental results and are thought to be qualitatively correct even when the assumption of continuous variation is no longer satisfied.

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* This article arises from work on semi-conductors that is being undertaken by the University of Reading, England, in behalf of Standard Telecommunication Laboratories Limited.

[1] All numbered references will be found on p. 177.

For a single homogeneous crystal, σ_0 should be proportional to the square root of the concentration of impurity centres.[24] For a micro-crystalline aggregate, this relation is generally not fulfilled, but it is found experimentally that some such correlation between σ_0 and the average concentration of impurity centres (as inferred from thermo-electric measurements) still exists. The correlation is not exact, indicating that the intergranular boundaries and the grains themselves make comparable contributions to the resistivity.

The thermo-electric properties of semi-conductors have a practical as well as a theoretical interest owing to the magnitude of the thermopowers involved. Thermo-couples consisting of a metal and a semi-conductor can have thermoelectric powers of the order of 1 millivolt per degree centigrade, which is between 10 and 80 times as high as those observed on purely metallic couples. Semi-conductors may thus be very suitable for the construction of sensitive thermopiles, and indeed have already been used for this purpose.

2. *Thermo-Electric Effects in Homogeneous Semi-Conductors*

To simplify the problem as much as possible, the following treatment concerns only semi-conductors that are either completely of the "excess" or completely of the "deficit" type, and that are not subject to "intrinsic" semi-conduction within the temperature range under discussion.

Fig. 1 shows the approximate energy relations at the contact between a metal and a typical excess semi-conductor. F is the Fermi-level of the metal, ϵ the activation energy of the semi-conductor, ϕ and χ are the thermionic work-functions of the metal and the semi-conductor, respectively, and V_D is called the diffusion potential. The potential barrier is of the type suggested by Schottky, but its precise shape or thickness is of no importance for present considerations. When contact is made, the relative heights of the Fermi-level and the conduction band adjust themselves so that the electrons on either side of the barrier are in dynamic equilibrium, i.e., so that equal numbers of electrons cross the barrier in opposite directions during any given time. This can only be if the electron concentration at any particular level, e.g., the peak of the barrier, is uniform. Since the electron concentrations decrease very nearly exponentially with increasing energy, the necessary equilibrium condition is

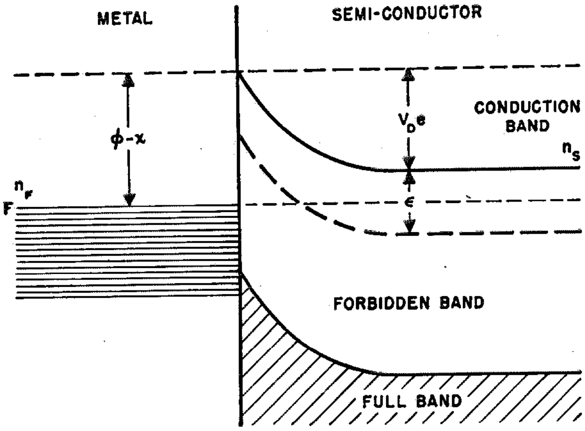

Fig. 1—Energy relations at the boundary between a metal and an excess semi-conductor.

$$n_F \exp\left[-(\phi-\chi)/kT\right]=n_S \exp\left[-eV_D/kT\right], \quad (1)$$

where n_F is the electron concentration at the Fermi-level of the metal and n_S that at the bottom of the conduction band in the semi-conductor. If Π is the contact potential at the junction, then

$$e\Pi=\phi-\chi-eV_D \quad (2)$$

and hence, after substitution into (1) and re-arrangement,

$$\Pi=(Tk/e)\log n_F/n_S. \tag{3}$$

This should not be confused with a similar equation sometimes encountered in the literature (and now known to be erroneous), in which n means the *total* electron concentration, independent of energy level.

As a first approximation, we can assume that n_S is given by some expression

$$n_S=A\exp[-\epsilon/2kT], \tag{4}$$

where the "concentration constant" A is nearly independent of temperature compared with the exponential term and is proportional to the square root of the concentration of impurity centres (compare (13)). Thus, we have

$$\Pi=(Tk/e)(\log n_F/A+\epsilon/2kT). \tag{5}$$

n_F is not strictly independent of temperature (it is expected to decrease slightly as T increases), but it is convenient to neglect this for the present, i.e., to assume that the Thomson coefficient of the metal is negligible compared with that of the semi-conductor. Integration of (5) gives

$$E=\int_{T_2}^{T_1}(\Pi/T)dT=(T_1-T_2)(k/e)\log n_F/A \\ +(\epsilon/2e)\log T_1/T_2, \tag{6A}$$

where E is in absolute units, or

$$E=(T_1-T_2)(300k/e)\log n_F/A \\ +(V_a/2)\log T_1/T_2,\ \text{Volts}, \tag{6B}$$

where V_a is the activation potential in volts. The polarity of the electromotive force is such that the metal electrode at the hot junction becomes positive with respect to that at the cold junction. Equation (6B) already satisfies a number of important conditions. It is symmetrical in T, so that inter-change of T_1 and T_2 merely reverses the sign of the thermo-electromotive force. Similarly, $E=0$ when $T_1=T_2$. Inspection also shows that the law of intermediate metals and the corresponding law of intermediate semi-conductors are obeyed.

The thermo-dielectric power is, of course, given by

$$(dE/dT)_T=(300k/e)\log n_F/A+V_a/2T, \tag{7}$$

according to which dE/dT cannot be zero at any temperature, unless $n_F<A$. Cases in which this condition is definitely satisfied are not known.

Re-arrangement of (6B) leads to

$$\frac{E}{T_1-T_2}=\frac{300k}{e}\log\frac{n_F}{A}+\frac{V_a}{2}\left(\frac{\log T_1/T_2}{T_1-T_2}\right). \tag{8}$$

Thus, if the present considerations are correct, a straight line should be obtained by plotting the average thermo-power $E/(T_1-T_2)$ against the function $(\log T_1/T_2)/(T_1-T_2)$. The activation energy should be directly deducible from the slope of this line, and its intercept on the ordinate should be a function of the concentration of impurity centres. The average thermo-power is expected to be high for semi-conductors of low impurity content, and vice versa.

The same argument is also applicable to a pure deficit semi-conductor, with the important difference that the resulting thermo-electromotive forces are of the opposite polarity, i.e., the metal electrode at the hot junction becomes negatively charged. It was previously thought possible to determine the character of a given semi-conductor on the basis of this criterion alone, but it will be shown that this is not always reliable, even when the sign of $\log n_F/A$ in (7) is known.

The present treatment is based on the assumption that n_F/A can be regarded as constant. A more accurate treatment would have to take account of the temperature dependence of A and of the finite Thomson coefficient of the metal. To do this with uncompromising precision would introduce considerable complications. It is, however, possible to introduce a second stage of approximation without great difficulty. To do this, it is convenient to put

$$n_F/A=(n_F'/A')T^q, \tag{9}$$

where q is some constant, and to remember that

$$\sigma_s-\sigma_m=\Pi/T-d\Pi/dT, \tag{10}$$

where σ_s and σ_m are the Thomson coefficients of the semi-conductor and the metal respectively. Substitution into the previous equations then leads to

$$\sigma_s-\sigma_m=\epsilon/2eT-qk/e. \tag{11}$$

Hence, if σ_s and σ_m were known independently, the value of the constant q could be estimated by this indirect method.

The Thomson coefficient of the semi-conductor can be simply calculated as follows. When two parts of a semi-conductor are at different temperatures, the electrons in the conduction band will reach equilibrium under the influence of two opposing tendencies: the normal process of thermal diffusion and the electric field resulting from any charge displacement. Electrons will drift from the hot to the cold regions, because their concentration as well as their kinetic energy increases very greatly with increasing temperature. When equilibrium is reached and no current flows, this tendency to drift must just be counter-balanced by the electric field. The condition for this is

$$ebnF_x = eD(dn/dx), \tag{12}$$

where b is the electronic mobility, F_x the local electric field, D the diffusion coefficient ($=kTb/e$), and n the *total* electron concentration in the conduction band, given by the well-known expression

$$n = CN^{\frac{1}{2}}T^{\frac{3}{4}} \exp\left[-\epsilon/2kT\right], \tag{13}$$

where C is a constant and N the concentration of impurity centres. We then have

$$\left.\begin{aligned} F_x &= (kT/en)(dn/dx) \\ &= (kT/en)(dn/dT)(dT/dx) \end{aligned}\right\} \tag{14}$$

or:

$$\begin{aligned}(kT/e)(\epsilon/2kT^2+3/4T)(dT/dx)& \\ = dE/dx = (dE/dT)(dT/dx).& \end{aligned} \tag{15}$$

But dE/dT is the Thomson coefficient* σ_s, so that

$$\sigma_s = \epsilon/2eT + 3k/4e. \tag{16}$$

Equation (13) is expected to hold with accuracy only as long as the concentration of free electrons is small compared with that of the impurity centres (see page 158 of reference 24), and (16) is, therefore, subject to the same limitation. It is necessary to state that this assumption may not be entirely justified in the case of titanium dioxide.

With regard to σ_m, the first approximation was zero, and as a second approximation $\sigma_m = 3k/2e$ is probably justified, although this is strictly correct only for a non-degenerate electron gas. We thus have

$$\begin{aligned}\sigma_m &= \sigma_s - \epsilon/2eT + qk/e \\ &= (k/e)(3/4+q) = 3k/2e,\end{aligned} \tag{17}$$

so that

$$q = 3/4. \tag{18}$$

This can now be substituted into (9) and hence into (5), resulting in

$$\begin{aligned}\Pi/T = (k/e) \log n'_F/A'& \\ +(3k/4e) \log T + \epsilon/2eT,&\end{aligned} \tag{19}$$

which can be integrated with respect to temperature to give in absolute units:

$$\begin{aligned}E = (T_1-T_2)(k/e) \log n'_F/A' + (\epsilon/2e) \log T_1/T_2& \\ +(3k/4e)\left[T(\log T-1)\right]_{T_2}^{T_1}.&\end{aligned} \tag{20}$$

After re-arrangement and conversion into volts, this becomes

$$\begin{aligned}\frac{E}{T_1-T_2} - 300\frac{3k}{4e}T_1\frac{\log T_1/T_2}{T_1-T_2} - 300\frac{3k}{4e}\log T_2& \\ = 300\frac{k}{e}\left(\log\frac{n'_F}{A'}-\frac{3}{4}\right)+\frac{V_\beta}{2}\left(\frac{\log T_1/T_2}{T_1-T_2}\right),&\end{aligned} \tag{21}$$

where V_β is again the activation potential in volts, corresponding to V_α in (6B). Its value can be deduced from the line obtained by plotting the left-hand side of (21), of which all quantities are known in terms of experimental results, against the usual $(\log T_1/T_2)/(T_1-T_2)$. Equation (21) should represent a better approximation to the facts than (6B), subject to the main assumption being valid, namely, that the semi-conductor is homogeneous.

3. *Thermo-Electric Effects in Heterogeneous Semi-Conductors*

The semi-conductors normally available for investigation are often not entirely homogeneous, and it is important to examine the anomalies that can occur as a result of heterogeneous structure. Two important types of heterogeneity can be envisaged. Thus, a semi-conductor may have a regular grain structure, but the concentration of impurity centres may change systematically from one side of the solid to the other. Alternatively, the substance may be a conglomerate of grains, which differ individually in impurity content in a random manner, so that numerous large concentration gradients are distributed throughout the solid. In practice, the two cases may, of course, arise simultaneously.

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* It was thought conceivable that thermal expansion might have a noticeable effect on the Thomson coefficient, in so far as it discourages electron diffusion away from the hot regions. The approximate calculation of this effect can proceed by the same method as before, taking into account that the electron concentration at any temperature T must be reduced by a factor $1/(1+cT)$, where c is the (volume) coefficient of thermal expansion. Using this correction, it follows that $\sigma_s = \epsilon/2eT + (k/e)[3/4 - cT/(1+cT)]$. Taking c as 10^{-5} centimetres per degree centigrade and T as 600 degrees absolute, substitution shows that the reduction of the Thomson coefficient by thermal expansion is negligible at all temperatures here considered.

Consider the effect of continuous concentration gradients on the Thomson coefficient of the semi-conductor. To substitute for n in (12), when N is a function of x, we must put

$$dn/dx = (\partial n/\partial T)(dT/dx) + (\partial n/\partial N)(dN/dx), \quad (22)$$

which gives

$$\frac{dn}{dx} = n_x\left(\frac{\epsilon}{2kT_x^2} + \frac{3}{4T_x}\right)\frac{dT}{dx} + \frac{n_x}{2N_x}\cdot\frac{dN}{dx} \quad (23)$$

and, after substitution into (14), we obtain

$$\left(\frac{dE}{dT}\right)_x = \sigma_x = \left(\frac{\epsilon}{2eT_x} + \frac{3k}{4e}\right) + \frac{kT_x}{2eN_x}\frac{(dN/dx)}{(dT/dx)}. \quad (24)$$

It will be noted that this reduces to the previous form (16) when dN/dx is zero. Otherwise, the local value of the Thomson coefficient depends on the ratio of the concentration gradient to the temperature gradient.

If the specimen is entirely at a uniform temperature T, then the first term in (24) cannot make any contribution to the voltage, since its temperature integral between identical limits is zero. The second term, however, gives rise to an electromotive force E_0 between any two points that are characterised by impurity concentrations N_1 and N_2 respectively given by

$$E_0 = (kT/2e)\int_{N_2}^{N_1} (1/N)dN$$
$$= (kT/2e)\log N_1/N_2. \quad (25)$$

This electromotive force may be termed a "distributed contact potential." The influence of this distributed contact potential on the conventional representation of energy levels is shown in Fig. 2. The voltage E_0 is not directly measurable by means of contact electrodes, because it is exactly balanced by equal and opposite contact potential at the two junctions.

If the junctions at the ends of a heterogeneous semi-conductor are *at the same temperature*, but temperature gradients exist elsewhere between the two contacts, then the contact potentials at the junctions and the distributed contact potential within the solid are not necessarily equal and opposite, and a residual electromotive force E_r can, therefore, appear between the electrodes, given by

$$E_r = \frac{k}{2e}\oint \frac{T_x}{N_x}\frac{(dN/dx)}{(dT/dx)}dT. \quad (26)$$

If dN/dx, as well as dT/dx, vary symmetrically between the two electrodes, E_r is zero, but if either variable is unsymmetrical, E_r will have a finite value, which may be of either sign.

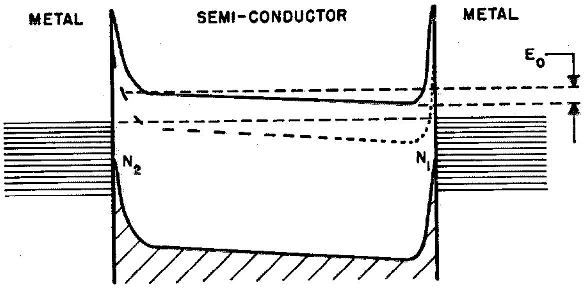

Fig. 2—Energy relations in an excess semi-conductor of non-uniform impurity content.

If the junctions at the ends of the semi-conductor are *not* at the same temperature, the normal thermo-electromotive force will, of course, appear in addition to the "distributed thermo-electromotive force" of magnitude

$$E_D = \frac{k}{2e}\int_{T_2}^{T_1} \frac{T_x}{N_x}\frac{(dN/dx)}{(dT/dx)}dT. \quad (27)$$

This electromotive force will have the greatest influence on the effective thermo-power when the temperature differences are small, and its presence should cause the apparent thermo-power to tend to infinity as $|T_1 - T_2| \to 0$. When the normal thermo-electromotive force is small, the sign of the effective thermo-power may be entirely dominated by E_D.

In the general case, the equation for the total thermo-electromotive force developed between two points at temperatures T_1 and T_2 and characterized by constants A_1' and A_2' becomes too complicated to be of practical use, namely,

$$\begin{aligned} E = (kT_1/e)\log(n_F'/A_1')T_1^{\frac{3}{4}} \\ -(kT_2/e)\log(n_F'/A_2')T_2^{\frac{3}{4}} \\ -(3k/4e)(T_1 - T_2) \\ +(\epsilon/2e)\log T_1/T_2 + E_D. \quad (28) \end{aligned}$$

If $A_1' \simeq A_2'$ and $|T_1 - T_2| \gg 0$, (21) should still be approximately obeyed.

These considerations show that as far as significant results can be obtained at all by experiments on heterogeneous materials, they necessitate the use of large temperature differences to minimize the effect of the distributed thermo-electromotive forces. In general, the properties of semi-conductors are very sensitive to changes of condition during their preparation. The existence of small variations of impurity content is, therefore, to be expected in sintered semi-conductors (e.g., as a result of uneven temperature distribution during their preparation) and probably in most others, particularly if the specimens are of considerable size.

The thermo-electric anomalies caused by the presence of concentration gradients are still expected to appear if the material concerned is a heterogeneous conglomerate of the second type, as envisaged above, provided that differences exist between the *average* impurity content of various regions in the solid.

4. Preparation and Micro-Structure of Titanium Dioxide Specimens

The specimens for the present investigation were prepared by pressing fine titanium dioxide powder into rectangular rods 10 centimetres long and 1 centimetre square under standard conditions. These rods were then pre-fired in air at about 1050 degrees centigrade for 1.5 hours. During this process, some volatile impurities are eliminated and the powder gains cohesion. The specimens were then slowly inserted into an electrically heated tube-furnace which was already at a temperature of 1000 degrees centigrade. During the four hours following, the temperature was raised to 1500 degrees and a mixture of 70 per cent nitrogen and 30 per cent hydrogen was passed through the tube. The specimens remained at the high temperature for four hours before the oven was allowed to cool. Each batch of four samples was then divided into two pairs. One pair was withdrawn from the oven after a cooling period of 25 minutes (1200 degrees) and then cooled further in the open air. These specimens will be referred to as "chilled," to distinguish them from those that were more "slowly cooled," i.e., that remained in the oven until its temperature had fallen to 700 degrees after 1.5 hours. These specimens were then further cooled while surrounded with asbestos packing. Owing to differential contraction, the number of breakages during "chilling" was considerable.

Chilled and slowly cooled specimens were expected to differ with regard to the concentration of ions which are "frozen" in interstitial positions, but no information was available as to whether the above differences in the rates of cooling are sufficient to have any appreciable effect in this respect.

The high-temperature sintering process in a reducing atmosphere results in a further elimination of volatile impurities and in a volume shrinkage of between 44 and 56 per cent. The reduced specimens are greyish-blue, with a metallic lustre, very hard and brittle. Their density is approximately 3.8 grams per cubic centimetre and does not vary between wide limits. The specimens are only slightly porous. X-ray powder-photographs have shown that they consist of rutile, and there is no evidence of the existence of any second crystalline phase.

The specimens were examined under the microscope in the form of very thin layers (20–30 microns) and as a very fine powder (capable of passing through a No. 100 mesh), both mounted in Canada balsam. It was found that these sintered bodies are not of homogeneous composition, but consist of four types of grains.

A. Small, faintly yellow grains, slightly bi-refringent and slightly pleocroic, having the characteristic appearance of normal rutile.

B. Grains of similar appearance except as regards their colour, which varies from slightly grey to dark blue-grey.

C. Fairly large black grains, often needle shaped, interspersed amongst a continuum of types *A* and *B*, and much more frequent in regions where the latter type is common.

D. Foreign impurities of two kinds: small grains of a red crystalline material, thought to be haematite, and somewhat larger grains of a material whose refractive index is very nearly the same as that of Canada balsam, probably quartz. Both kinds are only present in minute proportions and could not make any measurable contribution to the conductivity.

It was later found that the concentration of black needle-shaped crystals is greatest in the chilled specimens and smallest in specimens of low conductivity. The grain structure of the specimens is usually regular throughout the bulk of the material, but near the surface the proportion of yellow rutile grains increases sharply. This surface layer, which is of much higher resistivity than the rest of the material, is best developed on the slowly cooled specimens, i.e., those which have been for a longer time under conditions favourable for re-oxidation. The photo-micrograph, Fig. 3, shows the boundary between the bulk of the material on the left and the surface layer on the right. When a specimen is broken up into a fine powder, there is some tendency for grains of uniform colour (and hence, presumably, of uniform composition) to remain intact.

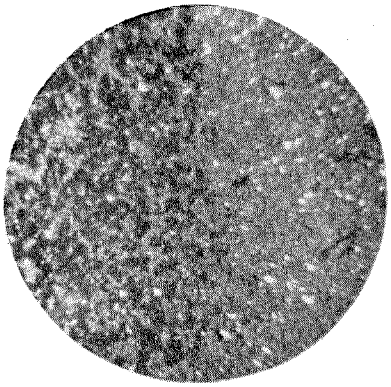

Fig. 3—Photo-micrograph of titanium dioxide specimen approximately 1.4 millimetres in diameter. Bulk material is at the left and the surface layer at the right.

The existence of this grain structure and the correlation of some of its characteristics with the method of preparation and the resistivity of the samples must inevitably be of great importance for the interpretation of the experimental results. There is reason to believe that these considerations are not confined to the present case, but are probably relevant to many other semi-conductors and to many previous measurements. It is important to realise that the normal theory of semi-conductors is not quantitatively applicable to conglomerates of the type here described.

5. *Experimental Technique*

As far as thermo-electric measurements on semi-conductors are concerned, either of two distinct methods can be adopted. One method consists of establishing a large temperature difference between two metallic junctions on the semi-conductor and measuring the resulting (total) thermo-electromotive force. The other consists of raising (or lowering) the whole specimen to the desired average temperature, and measuring the thermo-electromotive force due to a small temperature difference between the two junctions. If this temperature difference is sufficiently small, the measurement yields the true thermo-power at the temperature concerned. This is the method that has usually been adopted in the past, but it suffers from the grave draw-backs already discussed in Section 3. This is thought to be the reason for the low accuracy claimed by previous experimentors. Thus Morton's results[15] are consistent only within ±5 per cent, and those of Vogt[13] within even wider limits. On these grounds and also for reasons of greater experimental convenience, the first method was used during this investigation.

The simple arrangement is shown schematically in Fig. 4. The semi-conducting rod under test rested on two supports and was insulated from them by thin sheets of mica. On each side, a thin brass disc was firmly clamped to the specimen. The pressure was adjusted until the

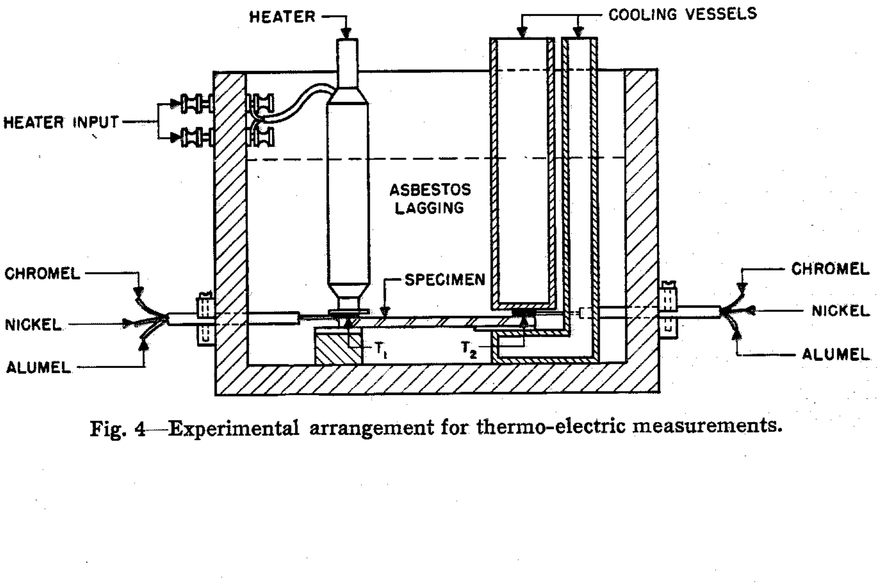

Fig. 4—Experimental arrangement for thermo-electric measurements.

contact resistance remained approximately constant. These brass discs, of which Fig. 5 is an enlarged representation, served as electrodes. Their tempertures could be determined by means of the attached "Chromel-Alumel" thermocouples. The thermo-electromotive forces otherwise measured are those between the semiconductor and nickel. One of the discs was heated by means of the heater, which was operated from a variable auto-transformer. To protect the associated galvanometer circuits in the event of a breakdown, the heater was electrically insulated from the electrode by a thin sheet of mica. The other electrode could be cooled from below and above by filling the vessels which form the clamp with solid carbon dioxide or liquid air. Cooling from below and above was found to be essential to make the best possible use of the cooling agent. The specimen was surrounded with asbestos lagging so as to exclude draughts, which were otherwise found to have a deleterious effect on the temperature stability of the system. To test whether slight oxidation of the electrodes could introduce substantial errors, some of the measurements were repeated, using gold-plated electrodes. Consistent results were obtained, and it can thus be assumed that if any oxide film is formed on the electrodes at temperatures below 450 degrees centigrade, it is too thin to introduce any appreciable temperature gradients.

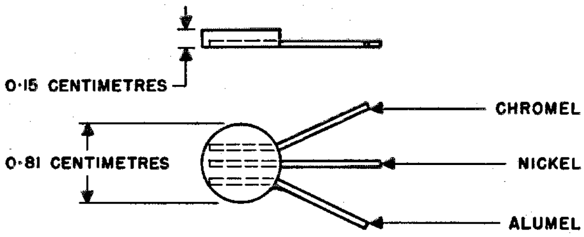

Fig. 5—Thermo-couple electrodes.

The whole arrangement is associated with a suitable potentiometer circuit. Each complete measurement thus demands the determination of T_1 and T_2 (by reference to oil-immersed cold junctions at a known temperature) and E, the resulting thermo-electromotive force between the semi-conductor and nickel.

The resistivity of the same specimens was measured between room temperature and 200 degrees centigrade, using the simple jig illustrated in Fig. 6. The potential probes, which were held in position by springs, rested on thin graphite lines which served to reduce the resistance in the probe circuit. The jig was mounted in an oven in turbulent air, and the potential difference between the probes measured at various temperatures for a constant current of 10 milliamperes.

It soon became evident that special precautions must be taken to eliminate unwanted thermo-electromotive forces in the potential probe circuit, or to correct for their presence. In view of the large thermo-powers developed by the junctions of the semi-conductor and the probes, very small temperature gradients along the specimen can lead to substantial errors in the determined resistivities. As small temperature gradients could not be entirely eliminated, a correction was made by measuring the themo-electromotive force between the potential probes when the specimen was disconnected from the battery circuit. This procedure greatly improved the consistency of the results.

6. *Resistivity Measurements*

The room-temperature (20 degrees centigrade) resistivity of the present specimens lies between 0.10 and 0.18 ohm–centimetres and is independent of the applied voltage. The temperature coefficient at 20 degrees centigrade is of the order of 1 per cent per degree centigrade and decreases with increasing temperature. All resistivity measurements between room temperature and 200 degrees centigrade were found to be strictly repeatable. Fig. 7 shows that—as might be expected—there is some correlation between the resistivity at room temperature and the percentage shrinkage during preparation. (All resistivities quoted are "effective" values. The

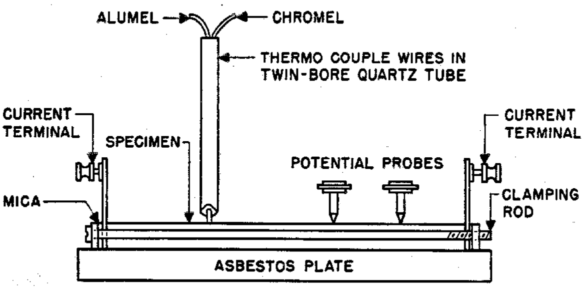

Fig. 6—Experimental arrangement for resistivity measurements.

existence of the more highly resistive surface layer of the specimens was neglected in the calculation.)

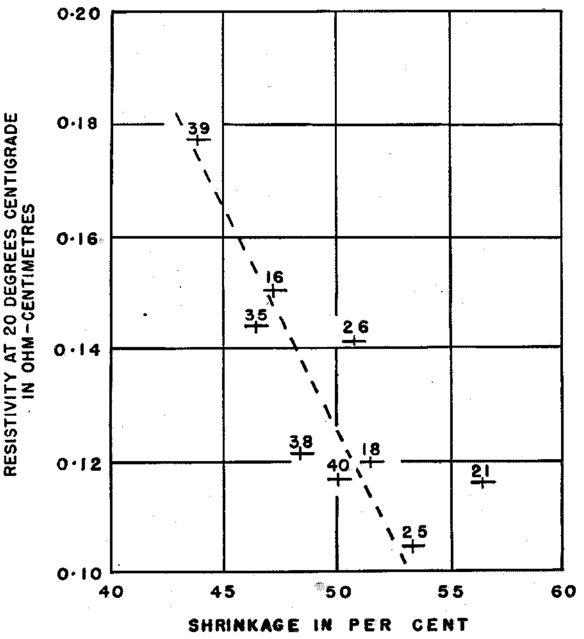

Fig. 7—Relation between resistivity at room temperature and shrinkage of specimens during firing. The specimen numbers are quoted.

No significant difference between chilled and slowly cooled specimens was noted. This seems to indicate that the process of chilling was not sufficiently drastic to make any appreciable difference in the concentration of ions frozen in interstitial positions. Alternatively, it may be that the critical temperature below which the presence of interstitial ions affects the resistivity measurements is considerably lower than room-temperature. (See page 160 of reference 24.)

To obtain the effective activation energies of the specimens, the results can be plotted in two alternative ways. For small temperature ranges, the expression

$$R = R_0 \exp [eV_a/2kT] \qquad (29)$$

is often a sufficiently accurate representation of the experimental results. A value for V_a can then be deduced from the slope of the line obtained by plotting R against $1/T$. However, for measurements over the temperature range here employed, these lines are all noticeably curved, so that only average gradients can be found from them. This means that the temperature dependence of R_0 is not negligible compared with that of the exponential term. This does not seem to have been observed during some previous experiments on titanium dioxide.[25] An expression for R, which is theoretically more satisfactory than (29), is

$$R = KT^{\frac{3}{4}} \exp [eV_b/2kT], \qquad (30)$$

where K is a constant. A straight line should thus be obtained by plotting $\log R/T^{\frac{3}{4}}$ against $1/T$. This was found to be the case, as shown in Fig. 8 for three typical specimens.†

A survey and summary of the results will be found in Table I. Wherever applicable, intercepts have been corrected for the different cross-sectional areas of the various specimens. It will be noted that V_a and V_b, the two activation potentials deduced from (29) and (30) respectively are not identical. V_b must be regarded as the more correct and significant value, even if its fundamental meaning is still somewhat in doubt.

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† For convenience, a multiple of the voltage V_p between the potential probes for a constant current (10 milliamperes) is used instead of the resistance R.

TABLE I

Specimen Number	Type	Resistivity at 20°C Ohm-cm	10^3K Ohm-cm	V_a Volts	V_b Volts	V_α Volts	V_β Volts	$\frac{3k}{e} \log \frac{n_F}{A} \times 10^5$ $mV/°C$
			(30)	(29)	(30)	(8)	(21)	(8)
16	chilled	0.151	0.179	0.081	0.127	0.033	0.076	0.248
18	chilled	0.120	0.114	0.088	0.139	0.060	0.116	0.165
21	chilled	0.116	0.082	0.078	0.127	0.060	0.122	0.193
25	slow cooled	0.105	0.133	0.076	0.129	0.046	0.104	0.191
26	slow cooled	0.141	0.176	0.072	0.125	0.037	0.089	0.207
35	slow cooled	0.144	0.202	0.071	0.117	0.036	0.088	0.275
38	slow cooled	0.122	0.073	0.118	0.168	0.014	0.076	0.134
39	slow cooled	0.177	0.123	0.107	0.152	0.025	0.084	0.183
40	slow cooled	0.117	0.077	0.108	0.166	0.029	0.084	0.162

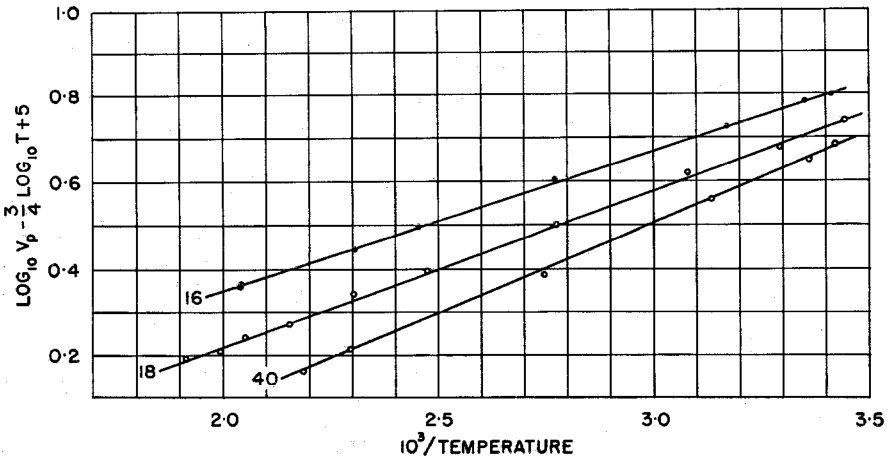

Fig. 8—Effect of temperature on the resistance of three specimens. $R=KT^{\frac{3}{4}} \exp [eV_b/2kT]$. V_p is the probe voltage and is proportional to R.

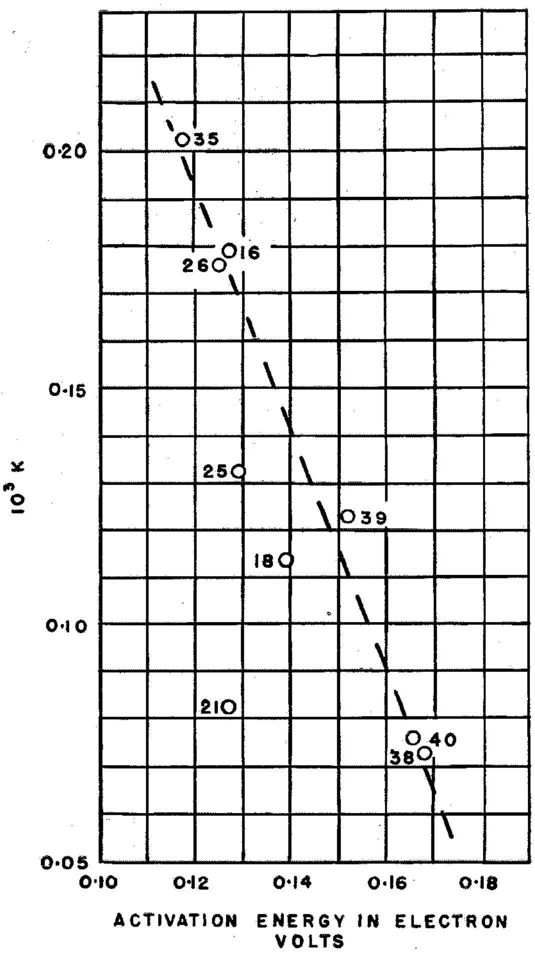

Fig. 9—Correlation between experimental values of concentration constant K and effective activation energy (V_b).

Meyer and Neldel[26] have proved the existence of a definite relation between V_a and R_0, contrary to earlier theoretical expectations. Fig. 9 shows that some such correlation also exists in the present case. (V_b and K are actually plotted.) The values obtained for V_a fall within narrow limits compared with those quoted by the above authors. (From 0.049 to 1.044 volts.)

7. *Thermo-Electric Measurements*

Measurements of the thermo-electromotive force of titanium dioxide–nickel couples were carried out over a temperature range from −176 to +400 degrees centigrade. Fig. 10 gives the results of one such complete series, plotted in accordance with (21). The temperatures T_1 and T_2 are marked for representative points. It will be seen that over this wide temperature range the results are substantially in agreement with theory.[‡] In general, the larger the temperature difference between the two junctions, the closer is the agreement of the results with the predicted straight-line characteristic. The deviations which exist appear to be typical for the present specimens. Large temperature gradients in the neighbourhood of the cold junction result in slightly low values for the average thermo-power, whereas large gradients in the neighbourhood of the hot junction do the reverse. It is to be expected that the latter occur shortly after the heater current is switched on, the former mostly during the initial period of cooling by means of solid carbon dioxide or liquid air. As the cooling and heating processes continue, the temperature gradients become less sharp and, if T_1-T_2 is sufficiently large, their effect becomes negligible.

None of the points in Fig. 10 corresponds to a temperature difference of less than 70 degrees centigrade. If T_1-T_2 is appreciably smaller, large deviations from the predicted behaviour

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[‡] When the experimental results are plotted in accordance with (8), they likewise give a straight line, though of smaller slope.

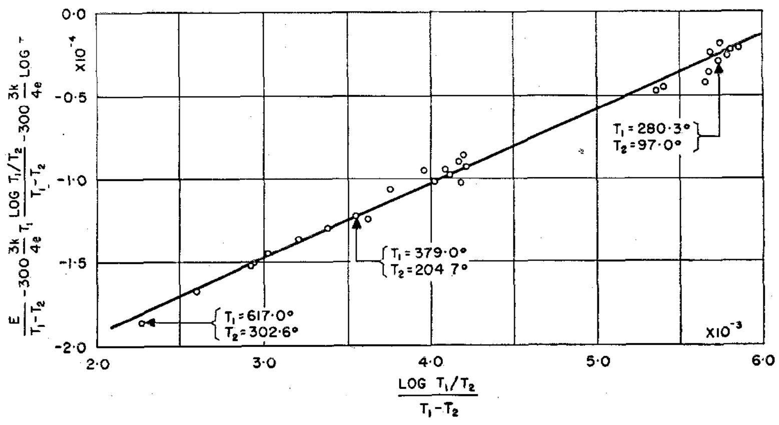

Fig. 10—Graphical determination of activation energy from thermo-electric measurements. Plot of experimental results in accordance with (21). Temperatures T_1 and T_2 are indicated for representative points. Specimen 26.

are observed. These deviations are interpreted as the direct consequence of the distributed thermo-electromotive forces that are expected to arise from the heterogeneous structure of the present specimens. The region of small temperature differences was further investigated in an attempt to isolate the normal thermo-electromotive force from the distributed thermo-electromotive force. To do this, a small asymmetrically wound heating coil was placed between the two main junctions (in the neighbourhood of the cold electrode) to produce the desired asymmetrical temperature distribution. Due to the presence of this subsidiary heater, T_2 rose to a constant value. The main heater was then switched on, and T_1 raised to and above T_2. The resulting thermo-electromotive forces were measured, and the results plotted in Fig. 11. They show clearly that the apparent average thermo-power tends to infinity as $|T_1-T_2|$ tends to zero, which means that a finite thermo-electromotive force E_r exists even when $T_1=T_2$. It will also be seen that the thermo-power of the couple is always positive provided $|T_1-T_2|$ is large, but can be of either sign when $|T_1-T_2|$ is small. If the subsidiary heater is displaced, the relative positions of the temperature and concentration gradients are also changed and this can be shown to have a considerable effect on the magnitude of E_r. The pairs of lines marked *a*, *b*, and *c* on Fig. 11 refer to three positions of the subsidiary heater. It will be noted that E_r changes sign between positions *a* and *b*, which seems to indicate that the average impurity content varies approximately symmetrically along the sample. Under the conditions to which Fig. 11 refers, E_r is +5.9 millivolts for position *a*, +1.5 millivolts for position *b*, and −6.4 millivolts for position *c*. There is, of course, some position between *a* and *b* for which E_r is zero.

Thermo-electromotive forces of this type have already been observed in the case of metals by

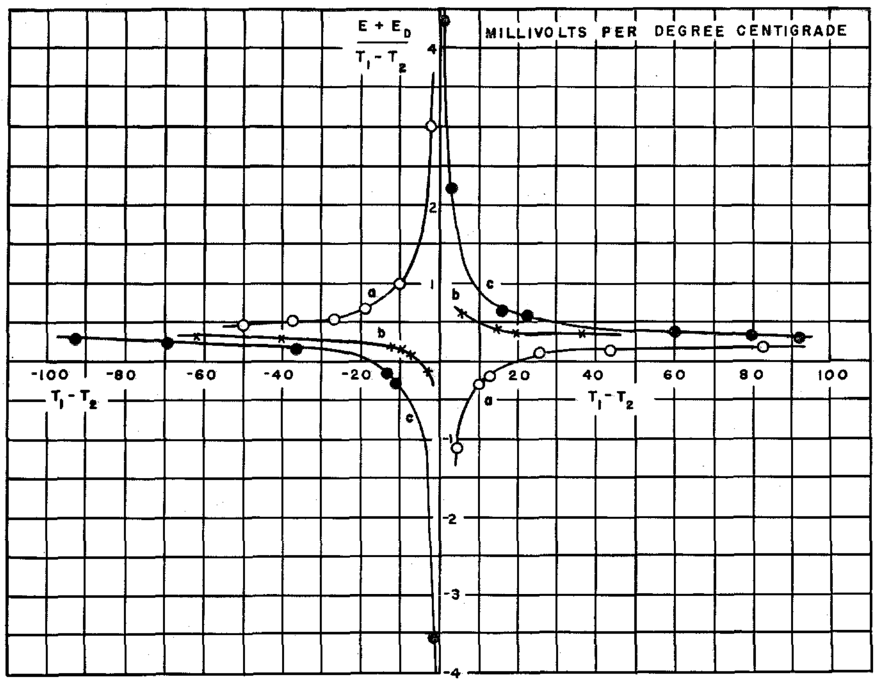

Fig. 11—Effect of distributed thermo-electromotive forces on apparent average thermo-power of the heterogeneous specimen 36.

Feng and Band,[27] though these authors did not associate them with any heterogeneous structure of the materials used. The distributed thermo-electromotive forces observed on the present samples are higher than those of metals by about two orders of magnitude.

The sign of the thermo-power of a semi-conductor–metal couple is generally taken as evidence for the character of the semi-conductor.[19] The demonstrated existence of distributed thermo-electromotive forces means, however, that this criterion is not valid unless the semi-conductor is known to be entirely homogeneous or unless the measurements are carried out for large temperature differences. In the general case, there is a further ambiguity owing to the fact that $\log n_F/A$ in (7) can be of either sign. Hence, it is important to know not only the sign of the thermo-power at a given temperature, but the exact characteristics of the couple over the widest possible temperature range. Conclusions derived from single measurements using small temperature differences can be entirely misleading in this respect.

Values for the activation potentials V_α and V_β were deduced from the experimental results and incorporated in Table I. It will be noted that V_b and V_β are not identical, but agree as regards general order of magnitude. Curiously enough, the agreement is best in the cases of specimens 21, 18, and 25, i.e., those which on Fig. 9 differ most prominently from the remaining samples. Otherwise, we have $V_\alpha < V_a$ and $V_\beta < V_b$, without any significant correlation between these quantities.

8. *Discussion*

Two important problems arise in connection with the present—and indeed many previous—measurements. Firstly, it will be observed that the nine specimens on which measurements have been carried out have yielded 15 different values for the activation energy. There is, in fact, nothing to suggest that V_β and V_b cannot vary continuously over a wide range of voltages. If eV_b (or eV_β) is the width of the energy gap between the impurity level and the conduction band of the solid, these results would mean that the width of this gap can vary continuously. This is very hard to envisage. On the contrary, it is to be expected that there are only a limited number of positions in the unit cell of the crystal lattice in which an interstitial atom can be accommodated, and while each of these may be associated with a different activation energy, the total number of such values should be small. Secondly, there is no *a priori* reason for any correlation between the activation energies and the impurity contents of various specimens, unless, of course, the impurity content is so large that the centres interfere with one another. This interference, although possible, does not seem to be the general solution of the problem, since Meyer[25] has shown the existence of this correlation even for specimens which were excellent insulators at room temperature, i.e., specimens in which the concentration of impurity centres must have been very small indeed.

Since the present specimens are not homogeneous, but contain grains of different impurity content, it appears reasonable to associate the multiplicity of values for V_a and V_b (similarly V_α and V_β) with the multiplicity of proportions in which grains of different purity may constitute any particular specimen. V_a and V_b then have the significance of "effective" activation energies, and are dependent on the activation energies of the different types of grains. The problem in its simplest form can be formulated as follows: What are the effective values of R_0 and V_a of a solid conglomerate composed of semi-conducting grains whose individual constants or group constants are $r_1, r_2, \ldots., r_n$, and $V_{a1}, V_{a2}, \ldots., V_{an}$ (so that $R_n = r_n \exp [eV_{an}/2kT]$), and which are present in concentrations $c_1, c_2, \ldots., c_n$, respectively? It is assumed that r_n depends only on the concentration of impurity centres within each type of grain and can, therefore, vary continuously, whereas the activation energies V_{an} are limited to a small number of definite values. To obtain an approximate solution, Borelius and co-workers[28] have suggested the application of a much earlier result due to Lichtenecker[29] on the resistivity of mixtures. Lichtenecker's calculations are mainly intended for two-dimensional models of two almost continuous constituents, but could be generalized in the following form.

The resistivity of the aggregate must be given by some function

$$R = F(R_1, R_2, \ldots, R_n;\quad c_1, c_2, \ldots, c_n). \qquad (31)$$

Lichtenecker also imposes the (arbitrary) condition:

$$1/R = F(1/R_1, 1/R_2, \ldots, 1/R_n;\quad c_1, c_2, \ldots, c_n). \qquad (32)$$

Further, we must have $R = R_n$ if $c_n = 1$ and $R = 0$ if $R_n = 0$ and $c_n \neq 0$. The simplest expression which satisfies these conditions is

$$R = \prod R_n^{c_n g_n}, \qquad (33)$$

where the quantities g are parameters that depend on the geometrical configuration of the grains. For dimensional reasons, we must also have $\Sigma c_n g_n = 1$.

Equation (33) cannot, of course, represent more than a first approximation. The treatment does not take account of the potential barriers between individual grains, nor does it seem probable that the characteristics of the configuration can be satisfactorily expressed by a single parameter for each constituent. Subject to these reservations, however, the expression is partially successful, in so far as it predicts the existence of an effective activation energy eV_a, where

$$V_a = \Sigma c_n g_n V_{an}, \qquad (34)$$

a quantity which must be expected to vary from sample to sample. Also, since

$$R_0 = \prod r_n^{c_n g_n}, \qquad (35)$$

some correlation between the values of V_a and R_0 is to be expected for specimens which have similar values of g.

If the resistivity of the specimens were entirely due to the properties of individual grains and not to structural factors, a close negative correlation should exist between resistivity and the concentration of impurity centres. If the present considerations are correct, the intercepts of the lines obtained by plotting the results of thermo-electric measurements in accordance with (8) should arrange the specimens approximately in the reversed order of their average impurity content. Fig. 12 shows that a correlation between these intercepts and those on Fig. 8 (representing the resistivities) does indeed exist. The fact that the correlation is not complete can be interpreted as meaning that the average impurity content of the grains and their geometrical configuration make comparable contributions to the resistivity.

The precise manner in which the different types of grain affect the resistivity of the substance as a whole remains a subject for future investigation. Because the grains presumably differ as regards their impurity content, the Schottky barriers at each grain contact represent an additional source of resistance. Each barrier of this type would normally have nonlinear and even slightly rectifying conduction characteristics, if the applied voltage is large. For small voltages, these resistances should be ohmic. The resistivity of the specimens as a whole was found to be strictly independent of the applied voltage, and it must therefore be assumed that the voltage developed across individual barriers is too small to produce nonlinear conduction, but that the total voltage across all barriers is an appreciable fraction of the total voltage applied.

The correlation of results shown in Fig. 12 confirms that semi-conductor–metal couples of high thermo-power are also expected to have a high resistance. This provides a natural limit of practical applicability, because high thermo-

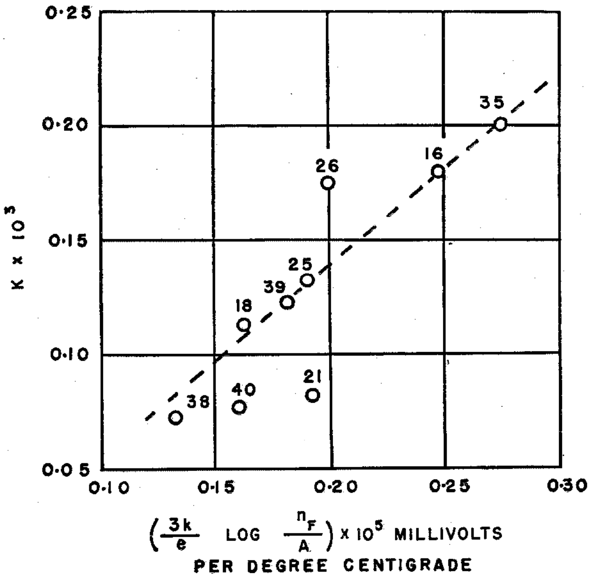

Fig. 12—Correlation between thermo-electric and resistivity measurements.

powers can only be usefully employed as long as the internal resistance of the couples is reasonably low.

With regard to the experimental values for the activation potential, agreement between V_b and V_β can only be expected as regards order of magnitude in the case of specimens that are not homogeneous. The internal mechanism by which activation energies can be "compounded" from a series of different contributions is still practically unexplored. There is no *a priori* reason to assume that the resultant value should be identical for conductive and thermo-electric processes. In the latter case, the heterogeneous structure is expected to result in "internal short-circuits" which would tend to diminish the total thermo-electromotive forces. This may account for the fact that V_b is always greater than V_β, but the problem does not seem to lend itself to more precise analysis. For homogeneous materials V_b and V_β should, however, be identical, and it would be of great interest to repeat the present experiments on substances that can be obtained in the form of single homogeneous crystals of suitable size.

9. *Summary of Main Conclusions*

A. A simple theory of thermo-electric effects, based on the electronic energy relations at a contact, leads to an expression (21) that is in qualitative agreement with experimental results over the temperature range −176 to +400 degrees centigrade.

B. The Thomson coefficient of a semi-conductor can be deduced from the equilibrium condition between thermal diffusion and electric fields. (Equations (16) and (24).)

C. Sintered titanium dioxide specimens consist of grains of different colour and hence, presumably, of different impurity content.

D. In a semi-conductor that is not homogeneous (i.e., in which the average concentration of impurity centres is subject to local variations), "distributed contact potentials" must exist throughout the solid, even when the temperature is uniform. Distributed contact potentials cannot be directly measured by means of contact electrodes, since they are equal and opposite to the normal contact potentials.

E. If temperature gradients exist in a heterogeneous semi-conductor, "distributed thermo-electromotive forces" appear in addition to the normal thermo-electromotive force. As a result, the apparent average thermo-power $(E_D+E)/(T_1-T_2)$ tends to infinity as the temperature difference between the junctions tends to zero.

F. Thermo-electric measurements can be used to grade a series of specimens in the order of their average impurity content. Comparison with the results of resistivity measurements then gives some guidance as to the extent to which the grain configuration affects the resistivity. The influence of configuration is detectable for the present specimens.

G. If only one mechanism of conduction is active, thermo-electric measurements can be used to determine the character of a semi-conductor (excess of deficit type), provided that large temperature differences are used and certain precautions observed.

H. The magnitude of the distributed thermo-electromotive forces gives some guidance on the homogeneity of specimens.

I. A correlation exists between the room-temperature conductivity of the specimens and the percentage shrinkage during preparation. The resistivity of the specimens is independent of the applied voltage.

J. Resistivity measurements between room temperature and +200 degrees centigrade are strictly repeatable. No differences have been observed between chilled and slowly cooled specimens.

K. In general, semi-conductors of high thermo-power have a high resistivity, and vice versa.

L. The temperature dependence of the resistivity between room temperature and 200 degrees centigrade is well expressed by the theoretical equation $R=KT^{\frac{3}{4}}$ $\exp[eV_b/2kT]$. In the case of a heterogeneous solid, the interpretation of the energy term eV_b is still in some doubt. It is thought probable that eV_b represents some form of effective activation energy, e.g., of the kind arising from Lichtenecker's equations.[29]

M. Rough agreement can be reached between the activation energies deduced from resistivity and thermo-electric measurements respectively. Exact agreement cannot be expected when the specimens are heterogeneous.

10. *Acknowledgments*

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Relationship Between Rate of Transmission of Information, Frequency Bandwidth, and Signal-to-Noise Ratio*

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THE history of communication is reviewed to show the gradual growth of modern theory, stressing in particular the major steps made by Carson with his sideband theory, and by Armstrong with his demonstration of improved quality of communication by use of increased bandwidth.

Established principles of communication are considered to show how they may be fitted together to form a coherent theory. After estimating the exact benefit to be gained by the use of the right balance between rate of transmission of information and the frequency bandwidth used in established expanded-band systems of communication, consideration is given to what particular features of transmission systems appear to be necessary for best performance.

Arising out of the above, a comparatively unknown system of modulation, tentatively called "step modulation," is shown to have all known characteristics of efficiency. The exact performance of the system is examined, and it is found that the process of bandwidth expansion to yield improved demodulated signal-to-noise ratio may be inverted to provide communication through reduced bandwidth at the cost of signal-to-noise ratio.

Finally, based upon the above analysis, a hypothesis is put forward to the effect that the new system may define the theoretical limit of efficient use of frequency bandwidth for any transmission system, for the case when the only known characteristic of the information wave is the frequency band that contains it.

1. *Object of Paper*

It is the object of this paper to produce an integrated modern theory of communication. At the present time, there appears to be considerable confusion concerning what use can be made of frequency bandwidth; various established principles do not appear to be co-ordinated into one general theory. Such a co-ordination should, it is believed, not only do much to establish a useful comparison between the efficiencies of well-known expanded-band systems of transmission, but may even permit us to specify a new ideal or maximum possible efficiency for the transmission of information according to the frequency bandwidth available.

The analysis does not include the conception of coding of specialised forms of information, but comprises only the problem of transmission of *any* waveform. No advantage will, for example, be taken of the fact that human speech usually carries much more information than the words that it comprises: we are here only concerned with transmitting any waveform with the maximum possible fidelity. Though the possibility of bandwidth compression arises, for the purpose of this study this must not be achieved by any coding process that might convert a voice showing particular personal characteristics, dialect, or "feeling" into a "mechanical" voice.

2. *Historical Review of Communication Theory*

The earliest electrical signalling was, of course, in the form of a code, information being carried in the form of a keyed direct current. Multichannel operation was later achieved by the use of keyed alternating currents of different frequency, and at this stage problems of bandwidth were already becoming significant.

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* A symposium on "Recent Advances in the Theory of Communication" was presented at the November 12, 1947, meeting of the New York Section of the Institute of Radio Engineers. Four papers were presented by A. G. Clavier, Federal Telecommunication Laboratories; B. D. Loughlin, Hazeltine Electronics Corporation; and J. R. Pierce and C. E. Shannon, both of Bell Telephone Laboratories. The paper by Mr. Clavier was entitled "Evaluation of Transmission Efficiency According to Hartley's Expression of Information Content" and has been submitted to the Institute for publication.

Electrical transmission of speech, which was made possible by translation of an acoustic wave by a microphone into a corresponding electric wave, marks an all-important advance in practical technique. It is interesting to note that before the days of radio, but following the use of equivalent telegraph systems, multiple speech transmission was visualised by the distributor or time-sharing method.

The modulation of a carrier wave by speech currents is probably the next major step, and this immediately brings us to the time when the communication world became conscious of a relationship between signal-to-noise ratio and selectivity. Sideband theory was not, of course, available; the necessary frequency bandwidth for the transmission of speech on a carrier wave was not appreciated, though strangely enough, the bandwidth requirements of telephone lines for transmission of the speech wave itself were well understood.

The development of sideband theory, and Carson's conception of the single-sideband method of transmission probably represent the greatest contribution to communication theory up to the present. The single-sideband method persisted for a very long time as the fundamental ideal of transmission, and to this day it represents the best standard of efficiency for making comparison between other systems.

Both before and after the conception of single-sideband transmission, a great deal of work was carried out on other lines in an attempt to suppress noise. All the methods developed, however, were based either on a fallacy or on special conditions such as the presence of a particular type of noise. For example, the use of a wide receiving bandwidth for amplitude-modulated transmission and the use of a limiter to remove very large but short peaks of noise are quite effective, and may give a greater signal-to-noise ratio than the use of the same power in single-sideband transmission. Such systems, however, as Carson himself pointed out, involve particular assumptions with regard to the prevailing noise.

The Carson theory that, when noise is completely random in nature, the signal-to-noise ratio of a transmission system depends entirely upon the power developed in the signal sidebands, held good for many years. In justice to Carson, it should be added that the theory still holds good in cases where final signal-to-noise ratio is rather low. When noise power, receivable in a bandwidth equal to the information-wave bandwidth, is equal to or greater than the receivable signal power, then single-sideband transmission stands as an ideal that cannot be surpassed in signal-to-noise performance by any system. In other words, if only ability to communicate is considered, then single-sideband working represents a standard of perfection; it is only high-quality communication that has caused a revision of fundamental theory or, rather, opened up a new field of possibility.

Up to the date of Carson's teaching, improvement in signal-to-noise performance was attempted by reduction of transmission bandwidth. Having been set a very convincing lower limit for bandwidth, no revolutionary improvement now appeared possible. Probably, the natural attempts to minimise bandwidth actually retarded the communication world in its next major step.

Armstrong appears to be the first to have established that in conditions of small or moderate, but not very great, noise of a completely random nature, the signal-to-noise performance of single-sideband transmission can be radically improved by the use of greater bandwidth. His convincing demonstration of the new standard of quality of transmission, which can be achieved by broad-band frequency modulation, represents an important advance in communication technique.

At about the same time, A. H. Reeves was apparently conscious of the same principle in theoretical studies of pulse-time modulation, but practical development was not so rapid as in the case of frequency modulation.

Since the outbreak of war in 1939, little has been written of communication theory. Doubtless the principles established by Carson and Armstrong are now well co-ordinated in the minds of most communication engineers, but the present paper attempts to establish clearly, and without appreciable mathematics, the significance of the fundamental parameters "sideband power content" and "frequency bandwidth." This rather simplified theory shows that well-known practical systems have very different performance characteristics, and we are prompted to wonder whether there can be an ideal system.

Expanded-band systems of well-known type all show unidirectional improvement of signal-to-noise performance with expansion of bandwidth from that which must be consumed by single-sideband transmission. The thought arises that if the parameter bandwidth can be expanded smoothly and usefully, it would appear inconsistent with nature to set an absolute lower limit to its value for this would probably represent a discontinuity of law. Now that we have established the fact that bandwidth expansion can improve signal-to-noise ratio, is it not conceivable that bandwidth compression can be achieved at the cost of signal-to-noise ratio? In the past, bandwidth compression has been sought partly, of course, to provide a greater speed of signalling in a given frequency band, but also in the belief that the effect of noise would be reduced or eliminated. A fresh examination of the problem, undertaken on the expectation that signal-to-noise ratio will suffer, should have a better chance of success.

The present paper shows that there is no obvious discontinuity in the expansion and reduction of bandwidth for transmission of a given rate of information by telegraphy, and after correlating telegraph and telephone practice, consideration is given to a little-known system, which may become of major importance. In 1938, A. H. Reeves proposed a new method for expanded-band transmission, which he called pulse-step modulation, in which the claim was made that speech may be transmitted entirely without noise at the expense of a constant but very small coding distortion.

This new system of transmission, which has recently received considerable attention by communication engineers and has been variously described as "pulse-code," "pulse-count," and "pulse-digital modulation," has not only a very high theoretical signal-to-noise performance, but also the unique feature of adaptability for both expansion and compression of transmission bandwidth. In the present paper, the suggestion is put forward that step modulation may represent a new and universal ideal system, in which maximum possible benefit can be obtained from the frequency bandwidth available. This conception would not displace the old ideal single-sideband transmission as the latter would remain not only as the particular case in which effective bandwidth available for transmission is equal to the bandwidth of the information wave, but also as the method for achieving a signal-to-noise ratio of unity with a minimum of signal power.

3. Early Principles of Communication

This section will be devoted to a short analysis of simple but important principles of communication that appear to have some significance in building up a co-ordinated theory. These principles are limited in useful application to those systems that were well known before expanded-band systems such as frequency modulation; they are, nevertheless, perfectly sound and need to be well understood before consideration of later developments.

3.1 KEYING SYSTEMS

3.1.1 *Keyed Carrier Wave*

In the simplest of all methods of transmitting information, the periodic interruption of a carrier wave, noise can always be made to have no effect by the use of a sufficiently small receiver bandwidth and reducing speed of keying until the receiver is able to follow it. The conclusion should not be drawn, as was very common in the early days of telegraphy, that noise can be eliminated by bandwidth reduction, for in this case we leave out the essential parameter of speed of signalling. The correct conclusion, which leaves out no essential parameter, is that for a given factor of safety in output signal-to-noise ratio, the speed of signalling is inherently limited by the amount of noise present in the transmission path.

3.1.2 *Multiple-Position Keying*

If the signal-to-noise voltage ratio of a keyed carrier system is great enough, then simple on-off or two-position keying may be replaced by three-position keying; i.e., on, half-on, and off keying. With sufficient signal-to-noise ratio, this principle may be extended to any desired number of key positions, hence permitting the use of a code that sends a vastly greater amount of information in a given time.

When noise is not a limiting factor, we may use either a code of greater complexity or key at a faster rate and, when noise does not permit a particular code and speed, either speed of transmission may be lowered or a simpler code may be used to provide error-free operation.

3.1.3 *Transmission of Infinitely Variable or Uncoded Waveforms*

The accuracy of transmission of an uncoded waveform can be improved indefinitely by reduction of bandwidth if the speed of sending is correspondingly reduced. Thus, if voice recorded on a gramophone is transmitted as modulation of a carrier wave, noise picked up in the transmission path will have a smaller effect and demodulated signal-to-noise ratio may be improved by reduction of playing speed and receiver bandwidth.

In the case of transmission of a "real" voice at its natural speed, a reduction in bandwidth will result in a corresponding loss in frequency fidelity, so that good signal-to-noise ratio is achieved by direct modulation only when noise conditions are suitable for transmission of a very complex step code of the type mentioned in the previous section.

3.2 SIDEBAND ANALYSIS OF TRANSMISSION SYSTEMS

All signals that are used for the transmission of information can be analysed to yield a system of sidebands, which correspond to the information wave, and other components which carry no information and can be considered as carrier waves. The best-known systems for transmission of speech are, perhaps, the amplitude-modulated carrier wave and the single-sideband system. The former contains a steady carrier wave unaffected by modulation and devoid of any "information" and a pair of equal sidebands that represent the information wave. Single-sideband transmission represents the same transmission with carrier and one sideband completely removed: alternatively, it can be considered simply as the original information wave displaced, positively or negatively, in frequency.

It is convenient to consider some systems as having a "complex" carrier wave, or a whole system of carrier waves. For example, an amplitude-modulated pulse train may be considered as a multiple-carrier amplitude-modulation system. The carrier wave in this case is a constantly modulated wave, which may be analysed as a whole system of "simple" carrier waves: modulation by the information wave produces a pair of sidebands corresponding to each of the simple carrier waves.

For efficiency of transmission, it is evident that the carrier wave or complex carrier wave can serve no useful purpose except the simplification of receiver circuits because, being constant, it could be manufactured and supplied at the receiver. Signal-to-noise ratio of a system using a carrier wave must, therefore, inherently suffer. In some carrier-wave systems, however, the penalty is not great except for very low levels of modulation as the carrier level is a function of modulation level and disappears entirely for the higher percentages of modulation: hence, peak signal-to-noise ratio is not affected.

Transmission systems could be grouped or classified in a number of ways, but the following is found convenient for the purpose of this paper.

A. Systems using sidebands only (e.g., single-sideband and other systems produced as the result of balanced modulation).

B. Systems having a constant carrier wave or a complex carrier wave, plus sidebands (e.g., amplitude-modulated carrier or amplitude-modulated pulse trains).

C. Systems having a single or a complex carrier that varies with modulation depth, plus sidebands (e.g., frequency-modulated carrier, and pulse-time modulation).

As a rule, systems of type *C* have a constant mean power of transmission, the growth of sidebands with modulation compensating exactly for decay of carrier level.

In all these systems, sideband power level rises linearly with modulation power level for very small values of modulation. In systems *A* and *B*, this is true for all values of modulation up to the maximum value, but in systems *C*, non-linearity commences fairly early without producing non-linearity of the overall transmitter-receiver system. Systems of type *C*, if modulated heavily, are outside the scope of the present section, their peculiar characteristics not being appreciated until after practical development of phase modulation.

3.3 SIGNAL-TO-NOISE RATIO FOR SYSTEMS OF LINEAR SIDEBAND VARIATION

3.3.1 *Single-Sideband Ideal*

The simplest linear transmission and reception system of perfect efficiency is the single-sideband method. The whole power of the signal is used for transmission of information, and the frequency bandwidth occupied is just the same as that of the original information wave. The mechanism of reception assures equal efficiency of demodulation and, hence, signal-to-noise performance for all signal levels.

Single-sideband performance represents the most satisfactory standard of efficiency for comparison of all other systems, and if conditions of random noise are such that single-sideband working yields a signal-to-noise ratio of less than unity, then *no* system can yield a better performance.

3.3.2 *Ideal Linear Receiver*

The following theoretical method of reception appears to represent an ideal method for all transmission systems of the linear type.

All transmission systems may be analysed to yield a system of sidebands. If at the receiver these sidebands are separately heterodyned, each yielding the original information wave plus noise, the separate demodulated outputs may be added together to yield a combined demodulated output. Such a receiver, which we shall call the "ideal linear receiver", may not be practical, but it is described here as a theoretical mechanism for the study of signal-to-noise ratio.

3.3.3 *Multiple-Sideband Systems*

It may be shown very easily that, applied to a transmission system of multiple sidebands of equal amplitude, the ideal linear receiver yields exactly the same demodulated signal-to-noise ratio that would be yielded by a single-sideband system of equal sideband power. The ideal linear receiver yields the maximum signal-to-noise ratio visualised by the early Carson theory, and still, of course, represents our theoretical ideal for all systems of linear sideband variation.

In a multiple-sideband system, maximum efficiency is not achieved unless all sidebands are of equal level, but if demodulated outputs are suitably weighted before combination, the imperfection factor is very small indeed.

The above conception suggests that high efficiency of a communication system depends upon uniform loading of the frequency band or bands occupied.

3.3.4 *Conclusions*

Single-sideband transmission gives a signal-to-noise performance that represents a theoretical ideal for all systems of linear sideband variation.

Signal-to-noise performance for any system is inherently imperfect according to the ratio of sideband content to total power radiated.

When a signal contains multiple sidebands, the demodulated signal-to-noise ratio does not necessarily suffer by this, but performance can never be improved by radiation of the same mean power as similar sidebands on different frequencies. (It appears from the above that signal-to-noise ratio cannot be improved by the use of expanded-band technique, but it will be shown in a later section that linear systems are, in effect, only equivalent to single-sideband working in true consumption of bandwidth.)

Before continuing to a consideration of non-linear modulation systems, something should, perhaps, be said to clarify the early conception that noise could be reduced by bandwidth reduction. Evidently, noise power picked up in the transmission path may be reduced by reduction of receiver bandwidth, but this is merely a question of adapting the receiver to the required transmission, and reduction below a certain limit must inherently leave out some of the required signal, causing either inability to receive the total information or a deterioration of signal-to-noise ratio. Though the single-sideband method of transmission appears to represent the system that uses the minimum possible bandwidth, it does not necessarily provide a better signal-to-noise ratio than any other system, for a multiple-sideband system of the linear type can give an equal performance.

4. *Principles of Expanded-Band Communication*

The term "expanded-band" is chosen deliberately to define a transmission system that inherently consumes more effective bandwidth than single-sideband transmission. Though the term "wideband" has sometimes been used, the latter is considered unsuitable, having been used also in the simple sense of channel-handling capacity.

4.1 GENERAL

The demonstration by Armstrong that continuous-wave phase or frequency modulation could, in some circumstances, provide a better signal-to-noise performance than single-sideband technique made a deep impression on the orthodox communication engineer, appearing at first to indicate the necessity for a revision of theory.

The essential difference between systems that are inherently limited to single-sideband performance and those that can give a greater demodulated signal-to-noise ratio has been identified as a difference that corresponds to linear or non-linear variation of sideband power with modulation power level. The reason for superiority of non-linear systems is not, however, easily apparent, and the practical engineer who always expects to have to pay for a radical improvement considers at once what "currency" he must use to "purchase" this radical benefit.

The answer, of course, is that the improved quality of transmission is obtained at the expense of additional bandwidth. Though, at first, additional bandwidth does not always appear to provide improved signal-to-noise ratio, succeeding pages will show that benefit may always be obtained from the *true* consumption of bandwidth.

Confining ourselves for the moment to a consideration of the phase-modulated continuous carrier wave, we see that for small amounts of modulation, bandwidth is constant, but that large modulation produces an increase in bandwidth. For large modulation, bandwidth is proportional to modulation voltage, and so, of course, is the output signal-to-noise ratio.

If we use "bandwidth" as an essential factor for estimation of signal-to-noise performance, we see at once that two simple and practical parameters only—"sideband power content" and "bandwidth"—are required. When modulation is small, signal-to-noise ratio is determined by sideband power content only, and for large modulation, sideband content is substantially constant at 100 percent, when bandwidth variation dominates. This method of analysis is simple and practical, and when applied to any continuous-wave system yields the result that signal-to-noise ratio is proportional to both sideband power content and bandwidth occupied.

4.2 EFFECTIVE BANDWIDTH OF A COMMUNICATION SYSTEM

Effective bandwidth of a communication system is here defined as the true bandwidth divided by the number of times this bandwidth may be used again by other transmissions of the same effective bandwidth without crosstalk between the various systems. Analysis of signal-to-noise performance of communication systems using this new parameter yields the definite result that performance is always dependent upon the effective bandwidth consumed.

In accounting for the performance characteristics of systems using intermittent signals, it is at first tempting to consider effective bandwidth as total bandwidth multiplied by the fraction of time utilised. Such a definition often yields the same result as the one given above but fails to account for the lack of improvement of a number of systems, for example, the amplitude-modulated carrier wave over the single-sideband system.

4.3 LINEAR SYSTEMS ARE NOT "EFFECTIVE WIDEBAND" SYSTEMS

It can be shown that all systems of linear sideband variation do not effectively consume more bandwidth than the single-sideband method, and as a corollary this means that signal-to-noise performance cannot exceed that of single-sideband working.

It will be noticed that two separate double-sideband systems may be radiated on diphase carrier waves of the same frequency without mutual interference, so that the effective bandwidth of a double-sideband system is not greater than for single-sideband transmission. Taking a more general case, if two sidebands a are radiated on different frequencies in any proportions defined by k_1a and k_2a, then crosstalk does not need to be experienced from another communication channel using the same carrier frequencies, which radiates sidebands b in any different proportions (or phases) defined by c_1b and c_2b.

Separate demodulation of the frequency channels yields (k_1a+c_1b) and (k_2a+c_2b), when a combination in suitable proportions (and phases) can yield outputs containing either a without b, or b without a.

One special case of interest is the amplitude-modulated pulse train. If the information wave contains frequencies between zero and f cycles per second, then the minimum possible pulse-recurrence frequency is $2f$ per second. Now, video-frequency or direct-current pulses which are initiated in a total bandwidth of F cycles per second necessarily have a total duration of $1/F$ seconds, and if it is desired to interlace a number of similar channels with our considered channel, adjacent-channel time spacing without crosstalk may be equal to one half of the pulse width, i.e., $1/2F$ seconds. Such an arrangement would permit the use of F/f channels in the group period $1/2f$ seconds, so that effective bandwidth of a single channel is f cycles, the same as for single-sideband transmission.

Bandwidth consumption of a radio-frequency pulse train is not greater, despite the fact that a radio-frequency pulse of a given duration occupies twice the bandwidth of a video-frequency pulse, because two complete diphase systems can theoretically be radiated on the same carrier frequency.

4.4 SYSTEMS OF NON-LINEAR SIDEBAND VARIATION

A study of non-linear systems shows that effective bandwidth is always greater than for single-sideband transmission, and also that signal-to-noise ratio depends upon the value of this parameter. In many systems, signal-to-noise ratio is directly proportional to effective bandwidth, but the law of improvement is not always the same, as will be shown in the next section.

5. *Signal-To-Noise Ratio for Expanded-Band Modulation Systems*

5.1 GENERAL METHOD FOR DERIVATION OF PERFORMANCE CHARACTERISTICS

The performance characteristics of various expanded-band systems of transmission may be derived from a study of the particular receiving mechanism used for the system concerned. Obviously, the result obtained must be dependent upon the method used for demodulation, and as it is not always easy to prove that the method of reception must be ideal, it is felt that the method is not suitable for an academic study and comparison of various systems.

Another method has therefore been found, which derives the maximum possible performance figures for any system without reference to any practical receiving equipment. This method has been used to check figures derived by more practical logic, and wherever there is confidence that the practical receiving technique has no fundamental imperfection, agreement is perfect. The following processes are involved.

A. Suppression of modulation to a very small amount, when it may be considered that sideband production is linear with modulation, that effective bandwidth is not greater than for single-sideband transmission, and that sideband power level is less than noise power receivable in that bandwidth.

B. In the above conditions, it is not conceivable that any non-linear technique such as the use of limiters can assist in demodulation of the signal, and we accept Carson's definition of maximum possible signal-to-noise ratio, estimating this from the power of the sidebands present.

C. On increasing modulation level to correspond to practical expanded-band communication, signal-to-noise ratio must, of course, be raised linearly with modulation.

5.2 SIGNAL-TO-NOISE PERFORMANCE FOR CONTINUOUS-WAVE PHASE MODULATION

If we phase-modulate a carrier wave by a very small amount, then two sidebands only are produced, and the amplitude of each is $n/2$ times the amplitude of the carrier wave, where n is the deviation in radians.

$$\text{Total sideband power} = 2\times\left(\frac{n}{2}\right)^2$$

$$= \frac{n^2}{2} \text{ times carrier power.}$$

$$\text{Signal-to-noise ratio} = S_{\text{PM}}$$

$$= S_{\text{SSB}}\times\frac{n}{\sqrt{2}},$$

where S_{SSB} is the signal-to-noise ratio for single-sideband transmission when using the same mean power at the highest level of modulation.

But signal-to-noise ratio is always proportional to deviation, so the formula is correct for all values of deviation.

Thus, $S_{\mathrm{PM}} = S_{\mathrm{SSB}} \times \frac{n}{\sqrt{2}}$

$$= S_{\mathrm{SSB}} \times \frac{B}{b} \times \frac{1}{2\sqrt{2}} \text{ (for large modulation),}$$

where B = bandwidth and
b = bandwidth required for single-sideband working.

The same formula results also from a consideration of the performance of practical receiver circuits, hence confirming that an orthodox receiving technique represents an ideal receiving mechanism.

5.3 Performance of Pulse-Time Modulation

5.3.1 *Effective Bandwidth of a Pulse-Time-Modulation Channel*

We have shown, in Section 4.3, that the bandwidth consumed by an amplitude-modulated pulse train is the same as for single-sideband transmission. In this case, the time allotted per pulse was equal to the pulse rise or half pulse width.

In the present section, we shall consider only the case of pulse-time modulation of large time excursion, so that the time effectively consumed by a single pulse is equal to the total time that may be consumed by displacement modulation of the pulse. We shall neglect the comparatively small time consumed by the unmodulated pulse train.

With this approximation, it is evident that the effective bandwidth of a pulse-time-modulation channel is equal to

$$\frac{\text{total time excursion}}{\text{half pulse width}} \times \text{bandwidth for single sideband,}$$

5.3.2 *Signal-To-Noise Ratio for Pulse-Time Modulation (Video-Frequency Pulses)*

Let time of pulse rise $= t$ seconds, and pulse-recurrence frequency $= N$ per second. Frequency bandwidth $= 1/2t$ cycles per second. Number of harmonics of pulse-recurrence frequency present $= 1/2Nt$. Modulation *total* excursion of t seconds gives a phase excursion of pulse occurrence of $\pm\pi Nt$ radians. Phase excursion of top harmonic

$$= \pm\pi Nt \times \frac{1}{2Nt} = \pm\frac{\pi}{2} \text{ radians.}$$

Now reduce modulation $n\pi/2$ times, where n is very large, when the modulation of the top harmonic $= \pm 1/n$ radian. Each sideband of top harmonic $= 1/2n$ times harmonic amplitude.

Therefore, total sideband energy of top harmonic

$$= 2 \times 1/4n^2$$
$$= 1/2n^2 \text{ times energy of top harmonic.}$$

Now, sideband amplitude is proportional to phase excursion and, therefore, to harmonic order, so sideband energy is proportional to the square of harmonic order.

Sideband-power distribution is therefore parabolic with frequency, and total sideband power equals

$$\frac{1}{3} \times \frac{1}{2n^2} = \frac{1}{6n^2}$$

of total energy of transmission.

At this low level of modulation, where sideband power is linear with modulation power level, we cannot expect a better signal-to-noise ratio than that prescribed as a maximum by Carson.

Hence, signal-to-noise ratio for pulse-time modulation $= S_{\mathrm{SSB}} \times (1/6n^2)^{\frac{1}{2}}$ for modulation $2/n\pi$ of pulse rise (where S_{SSB} is the signal-to-noise ratio for single-sideband transmission, using the same mean power at full modulation)

$$= S_{\mathrm{SSB}} \times \left(\frac{1}{6n^2}\right)^{\frac{1}{2}} \times \frac{n\pi}{2} \times \frac{\text{total excursion}}{\text{time of pulse rise}}$$

$$= S_{\mathrm{SSB}} \times \frac{\pi}{2\sqrt{6}} \times \frac{\text{total time excursion}}{\text{time of pulse rise}}$$

$$= S_{\mathrm{SSB}} \times \frac{\pi}{2\sqrt{6}} \times \frac{\beta}{b} \text{ for wideband working,}$$

where β = effective bandwidth and
b = bandwidth required for single-sideband operation.

It may be noted that the constant factor $\pi/2\sqrt{6}$ differs from the corresponding factor of $1/2\sqrt{2}$ for phase modulation and, in fact, represents an improvement of performance of about 5 decibels. Unfortunately, in the practical case of pulse-time modulation applied to radio systems, actual bandwidth occupied by a multichannel system is twice the effective bandwidth, owing to the fact that the video-frequency system produces two sidebands on the carrier frequency used, so that practical performances of pulse-time modulation and continuous-wave phase modulation are very similar indeed, when using the same mean power of transmission.

5.3.3 *Special Note*

The theoretical method chosen for estimating signal-to-noise performance has found particular value in its application to pulse-time modulation. The equivalent formula developed from consideration of a practical receiver, in which the steep leading edges of pulses are used for demodulation, yields a result which represents an imperfection factor of about 2 decibels. The suggestion is made that such a receiver is inherently imperfect, and that the use of both edges of pulses, on which noise fluctuations are not identical, would assist in bringing performance very near to the theoretical ideal.

Alternatively, frequency-discriminator technique could yield a similar result.

5.4 PERFORMANCE OF TIME-SHARING FREQUENCY MODULATION

The present section represents a purely theoretical study of signal-frequency-modulated pulse trains, in which instantaneous values of the modulating signal are transmitted as signal-frequency values. Thus, a single communication channel can be considered as a frequency-modulated carrier wave that is "gated" into the form of short pulses. A number of such pulse trains, each bearing a different modulation, may be interlaced to provide a time-sharing multiplex system, and when the various pulses exactly fill up the whole of time, the total transmission is almost identical to that obtained by frequency modulating a continuous carrier wave by a multiplex amplitude-modulated video-frequency train. Thus, the multiplex system is conveniently described as time-sharing frequency modulation.

The apparent super-efficiency of signal-to-noise performance as compared with continuous-wave phase modulation or pulse-time modulation has not, so far as is known to the author, been confirmed by practical development, but there seems to be no reason for doubting it.

5.4.1 *Effective Bandwidth of a Signal-Frequency-Modulated Pulse Train*

In a signal-frequency-modulated pulse system, the pulses are not displaced in time by modulation, and the various pulses of a time-sharing multi-channel system may be permitted to overlap without theoretical crosstalk as were the amplitude-modulation pulses of Section 4.3.

Let the bandwidth occupied by a single radio-frequency pulse $=B_1$ cycles per second.

Let modulation frequency excursion $=\pm B_2$ cycles per second.

Then, total bandwidth $B=B_1+2B_2$.

Pulse rise $=1/B_1$ seconds.

For a modulating or information-wave bandwidth of b cycles, i.e., bandwidth for a single-sideband system, channel pulse intervals $=1/2b$ seconds.

$\therefore$ maximum possible number of channels $=B_1/2b$,

$\therefore$ effective bandwidth β per channel

$$=B\times\frac{2b}{B_1}=\frac{B_1+2B_2}{B_1}\times 2b.$$

Total bandwidth varies upwards with the amount of modulation, and in the special case when bandwidth increase due to modulation ($2B_2$) equals the total bandwidth of the unmodulated system (B_1), effective bandwidth per channel is $4b$, or four times that for single-sideband working.

5.4.2 *Signal-To-Noise Performance for Time-Sharing Frequency Modulation*

Maintaining the annotation of the previous paragraph, the deviation ratio for the highest modulation frequency b is equal to B_2/b.

In this state, all component carrier waves, which constitute the unmodulated repetitive pulse train, are modulated $\pm B_2/b$ radians, and if the amount of modulation is very small, only one pair of sidebands is produced for each carrier.

The power level of each pair of sidebands is $\frac{1}{2}\times\left(\frac{B_2}{b}\right)^2$ times that of the associated carrier power; therefore, total sideband power is $\frac{1}{2}\times\left(\frac{B_2}{b}\right)^2$ times total signal power. For a system in which deviation ratio is independent of modulation frequency, the signal-to-noise ratio would be

$$S_{\mathrm{SSB}}\times\frac{1}{\sqrt{2}}\left(\frac{B_2}{b}\right),$$

but in frequency modulation, deviation ratio is inversely proportional to modulation frequency. This has the effect of providing a triangular signal-voltage pattern or, alternatively, for correct fidelity, a tilted noise characteristic. Noise, therefore, has a square-law power distribution over the modulating-signal band, and total noise power is suppressed three times as compared with the system of constant phase deviation.

Hence, signal-to-noise ratio for time-sharing frequency modulation

$$=S_{\mathrm{TSFM}}$$

$$=S_{\mathrm{SSB}}\times\frac{\sqrt{3}}{\sqrt{2}}\times\frac{B_2}{b}$$

$$=S_{\mathrm{SSB}}\times\frac{\sqrt{3}}{\sqrt{2}}\times\frac{1}{b}\times\frac{B-B_1}{2}$$

$$=S_{\mathrm{SSB}}\times\frac{\sqrt{3}}{2\sqrt{2}}\times\frac{1}{b}\times(B-2nb),$$

where n = number of channels or

$$=S_{\mathrm{SSB}}\times\frac{\sqrt{3}}{2\sqrt{2}}\times\left(\frac{B}{b}-2n\right).$$

Thus, if n is small, performance approximates to that of continuous-wave frequency modulation, despite the fact that only a small part of time is occupied.

In terms of effective bandwidth β,

$$S_{\mathrm{TSFM}}=S_{\mathrm{SSB}}\times\frac{\sqrt{3}}{2\sqrt{2}}\times\left(\frac{\beta}{b}-2\right)\times n.$$

In the special case when modulation is sufficient to double overall bandwidth, when $\beta/b=4$,

$$S_{\mathrm{TSFM}}=S_{\mathrm{SSB}}\times\frac{\sqrt{3}}{\sqrt{2}}\times\frac{B}{4b}$$

or

$$S_{\mathrm{SSB}}\times\frac{\sqrt{3}}{\sqrt{2}}\times n$$

or

$$S_{\mathrm{SSB}}\times\frac{\sqrt{3}}{4\sqrt{2}}\times\frac{\beta}{b}\times n.$$

The last expression indicates an improvement over pulse-time modulation of $(n\sqrt{3})/4$ times, which may not have much significance for a 10-channel system, but would be of vast importance for a 200-channel system.

6. Noise Barrier

The mechanism of reception for any expanded-band system depends for correct operation upon the signal amplitude being greater than the noise amplitude in the transmission medium. For example, in the case of a time-modulated pulse train, peaks of noise equal to signal pulses would cause ambiguity in recognition of pulses, and immediate failure of the receiver. In a similar way, the mechanism for reception of a phase-modulated carrier wave depends upon the carrier amplitude exceeding the noise amplitude at the demodulator.

The effect of such breakdown is to introduce a new type of noise, the amount depending upon the frequency of breakdown.

6.1 PRACTICAL EFFECT OF NOISE BARRIER

In earlier sections of this paper, signal-to-noise performances have been estimated without reference to the breakdown point, which destroys the action of the receiving mechanism, and the impression might be gained that the formulae are quite accurate for moderate expansion of bandwidth, but that they are subject to a simple top-limit of bandwidth defined by peak noise. This, however, is not the case.

First, let us consider the characteristic of noise. For the purpose of the present study, we shall define noise as interference energy uniformly distributed over the frequency band, and which can usually be considered as a spectrum of equal but infinitesimal sine waves of all possible frequencies, spaced uniformly at infinitesimal intervals of frequency.

Such a definition comprises completely random noise, most natural static, and most forms of man-made static, but does not include what might be called "intelligent interference," comprising all interference energy to which the communication system is particularly susceptible, whether it be deliberately designed interference, many types of interfering signals, or some exceptional form of man-made static such as interference from a high-frequency furnace.

Under this definition of noise, though it is true that noise power is always proportional to bandwidth, the relationship between peak noise voltage (which determines the breakdown point) and root-mean-square noise voltage is not definable. In the case of completely random noise, the maximum noise peak in any bandwidth is theoretically infinite, and in all types of noise it is evident that very high peaks are possible.

In a practical communication system, occasional breakdown cannot be avoided, whatever the bandwidth chosen, but we may deduce the following.

In systems in which output signal-to-noise ratio is satisfactory but in which bandwidth expansion is small, the signal-to-noise ratio in the transmission medium is also satisfactory, so the frequency of breakdown is small, and the new source of noise has little significance. Bandwidth expansion can never reduce the frequency of breakdown, and in most cases will tend to increase it, so that useful expansion of bandwidth is limited. Thus, output signal-to-noise ratio expands with bandwidth, at first according to the law already defined by our formulae, then at a lower rate, usually passing through a maximum value and decreasing rapidly due to very frequent breakdown.

At first sight, consideration of the noise barrier appears to complicate our signal-to-noise calculations to a high degree, but this is hardly necessary in most practical applications of expanded-band systems, for the reason that such systems are normally designed to operate well within the bandwidth corresponding to the noise barrier. It is rare that bandwidth is so plentiful that it can be consumed without reference to the information-carrying ability of the communication channel.

As an example, let us consider the case of a multi-channel pulse-time-modulation system in which 20 separate speech channels are required, the demodulated signal-to-noise ratio of each channel being 60 decibels. The suggestion is made, here, that it would be impracticable to design equipment giving a signal-to-noise improvement over single-sideband working of 50 decibels, for the reason that total consumption of bandwidth would be excessive. It would appear that not more than about 20 decibels improvement over a single-sideband system could be designed for, when the theoretical bandwidth expansion factor would be about 20 for the video-frequency system or 40 for a high-frequency system, and the practical overall consumption of bandwidth is likely to be 80 times the theoretical bandwidth for single-sideband working. A practical bandwidth expansion of 80 times, on a 20-channel speech system of 3000-cycle-per-second fidelity, involves a transmission bandwidth of nearly 5 megacycles per second, which represents a practical receiver bandwidth. In this case, signal-to-noise ratio in the high-frequency transmission channel is about 40 decibels so that the frequency of breakdown due to high peaks of noise is quite insignificant.

In amplification of the above, it may be mentioned that for completely random or fluctuation noise the peak value of noise will exceed the root-mean-square value by a factor of 4.4 for only one part in 100,000 of time.

In cases where total consumption of bandwidth has little significance, the maximum useful expansion depends upon both the type of signal used and the characteristics of the prevailing noise. Significance of type of signal becomes evident when we consider that in a continuous-wave method of transmission the peak signal is independent of bandwidth, but it may be increased very much by the use of intermittent transmission. Thus, frequency modulation is much more susceptible to noise-barrier limitations than pulse-time modulation. The significance of the type of noise becomes evident when we consider that in the case of completely random noise the frequency at which the noise peak exceeds the root-mean-square value by a given factor is statistically definable and is independent of bandwidth. However, in the case of impulse noise, the peak to root-mean-square ratio is proportional to the square root of bandwidth, whereby harmless peaks in a narrow bandwidth become more and more significant with band expansion.

7. *Résumé of Established Principles*

A. Demodulated signal-to-noise ratio or overall quality of transmission corresponding to a given rate of transmission of information, is determined by the signal power available as sidebands and the frequency bandwidth available.

B. In general, increased bandwidth permits improved quality of transmission, but there is a limit to the maximum useful expansion.

C. Expansion of bandwidth towards the limit defined in *B* introduces a new source of noise, which eventually becomes the predominating factor, and expansion beyond this limit may involve a rapid deterioration of overall performance.

D. Maximum efficiency is obtained by the uniform energy loading of the frequency band occupied.

8. *Unification of Telegraph and Telephone Theory*

The analysis already given in Section 3.1 shows a very close connection between telegraph and telephone systems. A simple on-off telegraph system is able to transmit only a very simple waveform. The multiple-position keying system of Section 3.1.2 shows that when noise conditions are more favourable, a higher quality of transmission can be achieved by the use of a code of greater information-carrying ability. If the signal-to-noise ratio is large, then the number of possible positions or code units also becomes large, and a more complex sequence of units or a more complex waveform can be transmitted. In transmitting a speech waveform, then, we have in effect an infinite number of code units, and distortion by noise is inevitable.

It was a consideration such as the above which led A. H. Reeves to the conception of transmission of speech without any noise by his step-modulation principle. Before continuing to a description of this system, certain principles will be given that will help in understanding an extension of the simple telegraph to a telegraph system of multiple code units, and then to transmission of *any* waveform such as speech.

8.1 RATE AND QUALITY OF TRANSMISSION OF INFORMATION

8.1.1 *Maximum Rate of Signalling*

The maximum number of units of information per second that can be transmitted in a given bandwidth is equal to twice this bandwidth expressed in cycles per second.

This principle is clearly established from the conception of single-sideband transmission, for in the frequency band zero to f cycles per second, any combination of successive amplitudes up to a rate of $2f$ per second may be absolutely defined by a wave that contains frequency components up to f cycles per second maximum. In other words, a Fourier analysis of a waveform that gives an exactly correct result at instants uniformly spaced $1/2f$ seconds does not need to include frequency components higher than f cycles per second.

8.1.2 *Quality of Signalling*

The number of kinds of units of information, for example the number of different sizes of units, that can be transmitted and received without ambiguity depends upon the amount of noise prevailing.

8.1.3 *Production of Noise by Use of Complex Telegraph*

When the number of kinds of units to be transmitted exceeds the value that resolves all ambiguity, a breakdown of the system occurs causing an interference or noise voltage in the receiver, the transmitted waveform being reproduced with errors.

Transmission of smooth and infinitely variable waveforms such as speech involves the effective use of an infinite number of kinds of units of information and thus an inherent production of noise.

8.2 TELEPHONE TRANSMISSION WITHOUT NOISE

If a communication system is capable of producing an output wave in which instantaneous values lie between $\pm V$ volts, when noise voltage peaks cannot exceed $\pm(V/n)$ volts, then an unambiguous code of $(n+1)$ different units can be used. Thus, on transmitting a continuous waveform, this waveform could be first translated at the transmitter to a wave comprised of abrupt steps of amplitude having $(n+1)$ step levels, and circuit noise could never make any of the step levels ambiguous because the differential level of adjacent steps is twice the maximum noise voltage.

In receiving speech transmission by this method, it is not possible for circuit noise to modify the output waveform in any way, so that we can consider reception to be completely noise-free, at the expense of a small constant distortion in transmission. This represents the basis of the step-modulation principle proposed by A. H. Reeves, showing the final co-ordination between telegraph and telephone systems. This step system will now be considered in greater detail.

9. Step Modulation *

9.1 Suppression of Noise by Quantizing

The first principle of the system proposed by Reeves has already been expressed in the preceding paragraph. The steps must necessarily be imposed on the signal before transmission, rather than at the receiver, to avoid all ambiguity of receiver output due to noise. The receiver, being able to recognise all steps without ambiguity in a pre-arranged series of values in simple arithmetic progression, reproduces the same relative step amplitudes that were transmitted in the form of a series of instantaneous values that are nearest to the correct values of the original infinitely variable wave. Adjacent values are arranged to differ sufficiently to avoid ambiguity in the receiver through displacement by noise: that is, adjacent steps are separated by at least twice the peak noise voltage.

9.2 Coding of Signal Values

To the conception of quantizing of the information wave according to a convenient series of values, a process that may conveniently take place at a uniform sampling rate, Reeves added the principle of more accurate transmission of the waveform concerned by introducing a second channel of equal power, bandwidth, and sampling rate, this channel radiating a new series of step values, which may be used to provide a second-order approximation to the true instantaneous values of the information wave. For example, if the first channel transmits values corresponding to 0, 10, 20, 30, ..., 90, then the second channel transmits a similar series of values that denote discrepancies 0, 1, 2, 3, ..., 9 in the first series. The two channels are thus able to indicate all values between 0 and 99 without ambiguity due to interference by noise. Information-wave distortion may, of course, be reduced indefinitely by the use of additional channels.

At this stage, it became evident that the principle of sending accurate values by means of combinations of values represented a simple coding process; any desired overall accuracy of transmission could be achieved by using a suitable number of channels and step positions on each channel. In considering the maximum extension of bandwidth, as determined by the number of channels, Reeves' original patent makes the suggestions that speech might be transmitted with sufficient accuracy by means of 32 signal values, and that these could be produced by five channels, each of which has only two step positions. In other words, each channel works on a simple-telegraph on-off basis, the total number of combinations being 2^5 or 32.

Speech transmission by means of a binary code of on-off pulses may be achieved in a number of ways. Applications can be envisaged in which the five (or more) on-off channels might be applied to five different wire circuits. Alternatively, the various channels could be allotted to different frequency bands. Most likely, however, the time-sharing principle would be adopted for transmission of not only the code units of a single speech channel, but also the groups of units corresponding to many other speech channels, all in a single wideband channel.

9.3 Step Modulation Is Not a True Coding System

The step-modulation system has been described as a pulse coding system, as combinations of pulses are coded to mean different amplitudes of the signal wave. In this sense, of course, the process of transmission does involve a coding operation, but the coding operation is very different from the type of coding that has previously been proposed for the transmission of speech.

Just as a specialised letter code is only applicable to a particular language and even to a specialised field of interest, coding systems for speech, which have already been given that designation, have been dependent upon certain assumptions about the waveforms that could possibly be produced by the human vocal cords. The point to be made here is that the step system is capable of transmitting any kind of waveform, including all kinds of sound waveforms whether harmonious or otherwise.

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* The present paper was written in original form in October, 1946, when the author had no knowledge of any practical development of pulse-code modulation, as the system is known today.

Though the steps imposed upon the signal provide a semblance of coding of the type with which we have previously been familiar, the distortion produced may be controlled to any degree of precision required by the information.

In actual practice there is no necessity to tolerate the presence of any coding distortion. The author has suggested that if residual "true" noise is preferred to circuit distortion, this is achieved simply by radiating the channel that transmits the smallest signal steps, not in the form of a finite series of specific values, but in the form of the exact discrepancy of signal amplitude defined by the other channels. In this case, there can be no trace of a step in the demodulated signal, the circuit noise of the last or maximum-detail channel contributing the exact amount of signal required, plus a little noise just as in other systems of transmission.

10. Signal-To-Noise Performance of Step Modulation

Signals transmitted by step modulation can be considered as having no noise, but a constant circuit distortion, or as having a distortion noise level, the peak of which cannot be greater than half the difference between two adjacent step levels. As a rule, the type of distortion produced can be more easily tolerated than random noise of the same power level, so that if we treat the step distortion as noise, we shall appear to be making a conservative estimate of signal-to-noise performance. However, full advantage cannot be taken of the smaller nuisance value of step distortion, for if we use this to reduce the total number of signal values, during periods of low modulation level, percentage distortion is liable to rise by an intolerable amount. To avoid this defect, practical development has introduced the feature of logarithmic amplitude compression of the information wave before quantizing so that successive steps correspond to information-wave values arranged in geometric progression. For example, if a total of 32 signal values are used, 16 positive and 16 negative, these may be chosen at 1.5-decibel intervals, but converted to linear steps of approximately 6 per cent of maximum amplitude by amplitude compression. In such a system, the amount of distortion or step noise varies with signal output, but percentage distortion is almost independent of modulation level.

The present study will be confined to a consideration of coding of signal values arranged in simple arithmetic progression. In this case, though step noise is zero for the condition of no modulation, it is otherwise almost independent of modulation level, and comparison with other expanded-band systems is made easier.

First, however, to make a comparison of signal-to-noise performance upon an identical basis, we shall consider performance of the modified step system mentioned in the previous section. In this system, in which the maximum-detail channel is transmitted as an exact correction, noise has the same meaning as for normal transmission systems, and comparison is completely fair.

10.1 SIGNAL-TO-NOISE PERFORMANCE OF THE MODIFIED STEP SYSTEM

In the present analysis, it will be assumed that the various coding channels are each transmitted by single-sideband methods.

On dividing the available signal power into n equal parts for division among n separate channels, maximum channel signal-to-noise ratio $= S_{\mathrm{SSB}}/\sqrt{n}$.

This value represents the signal-to-noise ratio of the fine-detail channel. However, we have $(n-1)$ other channels, which may each comprise signal values corresponding to any of N information-wave values, the number of step positions chosen for each of these channels.

These $(n-1)$ channels have a total of N^{n-1} code positions, and these are effective in expanding the signal-to-noise ratio of the fine-detail channel by the same amount.

Hence, signal-to-noise ratio for the step system

$$= \frac{S_{\mathrm{SSB}} \times N^{n-1}}{\sqrt{n}}.$$

In this formula, n is the bandwidth expansion factor β/b used in analysis of other systems. The important feature to be noticed is that demodulated signal-to-noise ratio is not a simple function of bandwidth expansion as in the case of frequency and pulse-time modulations, but can be made very large if N can be large. The value of N, the number of step values, is of course limited by the noise-barrier conditions, and the exact limitation must depend upon the nature of prevailing noise.

10.2 CONSIDERATION OF NOISE BARRIER

The step system appears to be different from any other transmission system in that constants can be chosen that make the noise barrier or breakdown threshold a very serious consideration, even when only a very moderate bandwidth-expansion factor is used. In other words, the enormously improved suppression of the only type of noise that can have much practical significance in most transmission systems is attained only at the expense of reduced factor of safety against what has previously been considered as a breakdown effect. However, it is usually found that a design that has equal limitations imposed by two or more separate effects is more economical than one which accidentally avoids all but one of such limitations.

It is beyond the scope of the present study, to consider the exact optimum design for a step-modulation communication system, in which normal and breakdown noise impose equal limitations, but approximations will enable us to draw useful conclusions.

As previously stated, in the case of completely random noise, the peak value will exceed the root-mean-square value by a factor of 4.4 for only one part in 100,000 of time. For communication in which ultimate signal-to-noise ratio is moderate, the occasional breakdown caused by peak noise can be permitted, and we may assume an *effective* peak to root-mean-square ratio for noise that is not greatly in excess of 4.4.

In considering the performance of step modulation, this factor becomes significant in a determination of the maximum number of unambiguous values that can be transmitted in a single channel, and also in the number of coding channels that can be used, causing the noise barrier to assume a dominating importance despite the use of a small frequency band. We, therefore, introduce the factor p to define the ratio of effective peak to root-mean-square noise in a transmission bandwidth equal to that of the information wave.

10.3 VARIATION OF SIGNAL-TO-NOISE RATIO WITH BANDWIDTH-EXPANSION FACTOR

On splitting the available signal power into n equal parts for division among n separate channels, channel maximum signal to root-mean-square noise ratio $=S_{\mathrm{SSB}}/\sqrt{n}$ and maximum signal-to-*peak*-noise ratio $=S_{\mathrm{SSB}}/(p\sqrt{n})$.

In each channel, the available power may be used to provide N instantaneous values, this being achieved without ambiguity provided that

$$N<\left(\frac{S_{\mathrm{SSB}}}{p\sqrt{n}}+1\right).$$

The total number of combinations of the simultaneous n values $=N^n$, and the demodulated maximum signal-to-peak-noise ratio $S=N^n-1$.

The maximum value of this function corresponds to the maximum value of N, which is the integral value between

$$\frac{S_{\mathrm{SSB}}}{p\sqrt{n}} \quad \text{and} \quad \left(\frac{S_{\mathrm{SSB}}}{p\sqrt{n}}+1\right).$$

Therefore, maximum possible signal-to-*peak*-noise ratio S approaches very near to

$$\left(\frac{S_{\mathrm{SSB}}}{p\sqrt{n}}+1\right)^n-1.$$

This function is plotted in Fig. 1 for various values of S_{SSB}/p.

The expression is valid for both the normal step-modulation system, and also for the modified system in which the fine-detail channel is transmitted as an exact value. In the latter case, however, effective peak to root-mean-square noise is equal to p, so we can state that maximum possible signal to root-mean-square noise ratio is not greater than

$$p\times\left[\left(\frac{S_{\mathrm{SSB}}}{p\sqrt{n}}+1\right)^n-1\right].$$

In this expression, the number of channels n represents the bandwidth-expansion factor β/b, as used in earlier sections.

10.4 MAXIMUM SIGNAL-TO-NOISE RATIO FOR A GIVEN AVAILABLE POWER

The expression derived in the last paragraph appears, at first sight, to permit unlimited expansion of bandwidth and signal-to-noise ratio.

This, of course, is not true, the formula being valid only when channel signal-to-noise ratio is greater than unity. Thus $S_{SSB}/p\sqrt{n} > 1$.

Hence,

$$n_{\max} = \left(\frac{S_{SSB}}{p}\right)^2$$

and maximum signal-to-peak-noise ratio

$$S_{\max} = 2^{(S_{SSB}/p)^2} - 1$$
$$= 2^{n_{\max}} - 1.$$

The function $S_{\max}$ is plotted in Fig. 1 against $n_{\max}$ and this serves to show the maximum expansion of bandwidth possible, and the maximum possible signal-to-noise ratio obtainable for a given available power of transmission.

11. Bandwidth Compression

11.1 THEORETICAL POSSIBILITY OF BANDWIDTH COMPRESSION

The possibility of coding speech so that the normal information or meaning may be transmitted in a smaller bandwidth is very well established. Such methods, however, either leave out some characteristics of the voice, which may not be of importance in the sense that exact accent may not be required, or else the speech waveform may be assumed to have particular characteristics that are specially favoured by the transmission method chosen. This technique is outside the scope of the present study and cannot be considered, academically, as better than bandwidth reduction by reduction of frequency fidelity.

Early attempts at true bandwidth compression have usually been undertaken with the object of noise reduction, and this may be why no success has been achieved.

Earlier sections of this paper have established that signal-to-noise ratio may rise rapidly with increase in bandwidth, and also that with a given amount of power and bandwidth, signal-to-noise ratio can be raised by reduction of rate of transmission of information. Surely, then, if we increase rate of transmission of information in a given bandwidth, or if we maintain the rate of transmission in a reduced bandwidth, we must expect to reduce signal-to-noise ratio.

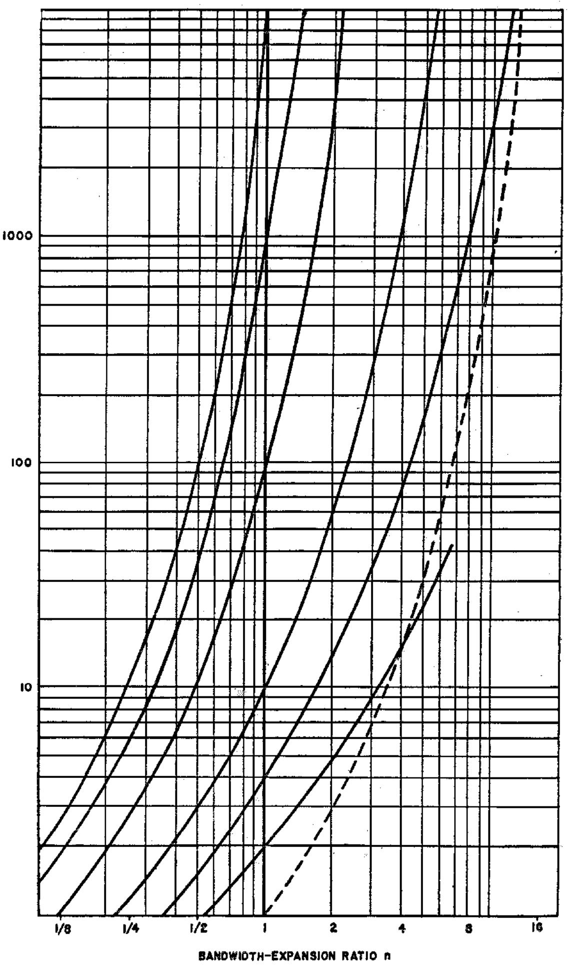

Fig. 1—Performance curves for step modulation.

$$S = \left(\frac{S_{SSB}}{p\sqrt{n}} + 1\right)^n - 1.$$

$$S_{\max} = 2^n - 1.$$

When $n=1$, $S = S_{SSB}/p$ as for single-sideband transmission.

It is evident from earlier sections of this paper that rate of transmission of information is not inherently limited by bandwidth. The amount of information that can be transmitted, when using a suitable code, depends not only upon the number of units of information per second (defined by bandwidth), but also upon the number of kinds of units, defined by signal-to-noise ratio. To transmit a waveform such as speech through a narrow band, we must evidently modify our speech waveform to have a smaller number of units per second, and to compensate for this, the quality or number of kinds of units must be increased.

This is the statement of principle expounded by the author in a discussion with A. H. Reeves in 1941, when the latter at once suggested, "Surely, then, all we need is the exact inverse of the step system," and proceeded to explain the exact mechanism for inversion.

Stated simply, if signal-to-noise ratio is such that by single-sideband transmission, 100 separate values can be transmitted without ambiguity, then these values could be coded to indicate all possible combinations of pairs of values of a set of 10 possible values. Thus, the rate of transmission of units may be halved if quality of unit is allowed to suffer. By this mechanism, bandwidth may be halved if signal-to-distortion ratio is allowed to fall from approximately 40 to about 20 decibels.

11.2 INVERTED STEP SYSTEM

An examination of Fig. 1 shows signal-to-noise curves extending on both sides of the single-sideband abscissa of bandwidth. This extension will be found to be perfectly valid if the horizontal scale is considered to be marked in effective bandwidth. In other words, if actual bandwidth is maintained constant and only half the pulse periodicity is used, effectively filling only half of time, then no assumptions are necessary for establishment of validity.

If it is desired actually to transmit a waveform in truly reduced overall bandwidth, we must pause to consider the effect of reduction of bandwidth on effective noise peak voltage.

For purely amplitude-modulated noise or impulse noise, the noise peak is proportional to bandwidth, and such an assumption would permit a better signal-to-noise ratio than is shown by the curves for bandwidth compression. For purely frequency-modulated noise, by which we mean noise concentrated in a very small bandwidth and very slowly sweeping over the frequency spectrum, then noise peak is substantially independent of bandwidth, and reduction of signal-to-noise ratio must be more serious than the curves indicate. For completely random noise, in which effective noise peak is proportional to the square root of bandwidth, the curves are valid for expression of signal-to-noise variation with true physical bandwidth.

The above conception does not, of course, deny the possibility of true bandwidth compression without assumptions with regard to noise, but a new set of curves would have to be produced, showing a rather more pessimistic performance for bandwidth compression. For the purpose of the present paper, we confine attention to effective-bandwidth reduction in which true bandwidth is not reduced, but the number of channels in a given bandwidth may be increased without theoretical limit.

The theoretical possibility of bandwidth compression is interesting from an academic standpoint, but it is very doubtful whether it can provide a useful engineering result. Quite apart from the difficulties involved in production of practical equipment, current problems would appear to favour a slight increase in bandwidth from the basic single-sideband system, rather than reduction, to provide satisfactory signal-to-noise ratio with economy of power.

12. Step Modulation as an Ideal System

Examination of signal-to-noise formulae developed for various communication systems suggests that there are two systems of outstanding efficiency of transmission, in their ability to use a given power and effective bandwidth to set up the greatest signal-to-noise ratio at the receiver output. These systems are the intermittent frequency-modulation system of Section 5.4 and the step-modulation system.

In terms of effective bandwidth, the intermittent frequency-modulation system appears to give the better result in some circumstances, but in terms of the same total bandwidth the step system is always superior. The frequency-modulation system appears superior (for constant effective bandwidth), if the prevailing noise is of the type in which effective peak to root-mean-square ratio is not increased by bandwidth expansion, as is the case for completely random noise. In the case of impulse noise, however, in which noise peak is proportional to bandwidth, the frequency-modulation system is very limited in expansion of total bandwidth.

Confining attention to the use of the same total bandwidth, it seems certain that the step system represents the most efficient system known, and in view of the fact that it appears to utilise every possible feature of communication efficiency, one is tempted to wonder whether it may represent an ideal that could never be surpassed. In support of this suggestion, one may list the following features.

A. The whole signal power is used to produce sidebands or information-bearing power.

B. Effective bandwidth is controllable for adaption to noise conditions.

C. Signal power uniformly loads the frequency band occupied.

D. Signal power uniformly loads the dimension time, hence minimising total bandwidth.

E. Given a fixed bandwidth, constants may be chosen so that normal noise and breakdown noise impose equal limitations.

F. It is interesting to consider the feature of reversibility of the process of bandwidth expansion and compression. It is likely that the most efficient system possible is capable of inversion, yet if so, there is only one reversible system. Inversion, without grave discontinuity, of a system that gives an inferior performance on one side of the single-sideband standard would give a better performance than the ideal on the other side of this abscissa. Thus, if the ideal system is reversible, there is only one reversible system, the step system.

In conclusion, though considerable evidence tends to set up the step-modulation system as a system of very high theoretical performance, it must not be assumed that it may eventually take the place of all others. Even if we confine attention to theoretical merit, the system does not dislodge single-sideband transmission from its position as a theoretical ideal, except when bandwidth available for transmission is in excess of single-sideband requirements. The single-sideband system can be considered as a special case of step modulation, in which the bandwidth consumed happens to equal the information-wave bandwidth. It may be that nature's choice of frequency bandwidth for the human voice is very fortunate, and that any highly developed communication system will find that this represents a natural optimum.

In considering practical difficulties, we may note that the single-sideband system was unable to dislodge the use of simple amplitude modulation from many applications, and that the considerable complexity of signal-coding devices may well retard or prevent the use of step modulation even in cases where large bandwidth is available. We can only say, with certainty, that in any case where available bandwidth is very large, theoretical performance of step modulation is vastly superior to that of single-sideband working.

13. Acknowledgment

The author wishes to acknowledge valuable criticism by Messrs. B. B. Jacobsen, K. G. Hodgson, V. J. Terry, and others, together with suggestions for expression of thought in a manner likely to achieve a wider range of appeal.

New Developments in Marine Radio Direction Finders*

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A REVIEW of the past and present state of marine direction finding and a comparison of its merit and fields of application with those of radar and loran are presented. Although marine direction finders have incorporated all of the design and component progress made in radio in the last 20 years, they are still not only working on the same basic principles, but the means of achieving measurements are practically identical to those used 20 years ago. Some of the reasons for this condition and factors that seem to limit further progress are reviewed.

A number of suggestions for improvement are: use of other parts of the frequency spectrum, use of pulse transmissions to reduce night effects, direct-reading indication of bearings, and combination of distance measurement with bearing measurement to obtain a fix from only one beacon station.

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It seems important to compare the marine direction finder with the new radio aids to navigation that were developed during the war, as a means of determining the causes for what seems to be an over-stabilization of design; and, as indicated by the title of this paper, to survey possible new developments in the field that may make the marine direction finder a still more useful instrument of navigation.

The technique of the simple loop direction finder operating on medium frequencies is now too well known to justify repetition in this paper. Also well known are propagation effects,[1] deviations caused by the ship's structure, and deviations due to other causes.

1. *Evolution and Present State of the Art*

The direction finder is the oldest radio aid to navigation. In the form of either the small rotatable loop or, before that, the large, fixed, crossed loops, the direction finder was the first instrument making use of the directive properties of electromagnetic waves to determine the direction of arrival or origin of a radio transmission. Early disclosures before World War I give the names of Stone, Bellini-Tosi, and Blondel as among the first to invent, propose, and try radio direction-finding systems. Later, Watson-Watt, Mesny, Adcock, Smith-Rose, and many others continued to establish the foundations of the new science by studying propagation and polarization effects, deviations, reflections, and other transmission vagaries.

Both fixed and mobile direction finders have been or can be designed to cover all of the frequency spectrum in which radio transmissions are made. Indications are very often obtained in a direct-reading manner, as through a mechanical pointer on a fixed scale (aircraft radio compass),[2] or in an instantaneous manner on the screen of a cathode-ray tube (fixed high-frequency direction finder).[3,4]

The marine direction finder for shipboard installation became practical with the availability of radio amplifiers after World War I. In the United States, the early work of F. A. Kolster, who in 1913 was already advocating the installation of shore radio beacons, resulted in a demonstration of the radio compass and position finder on board the U. S. Lighthouse Tender *Tulip* in

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* Presented before the International Merchant Marine Radio Aids to Navigation Conference, New London, Connecticut; April 30, 1947.

[1] H. Busignies, "Evaluation of Night Errors in Aircraft Direction Finding, 150–1500 Kilocycles," *Electrical Communication*, v. 23, pp. 42–62; March, 1946.

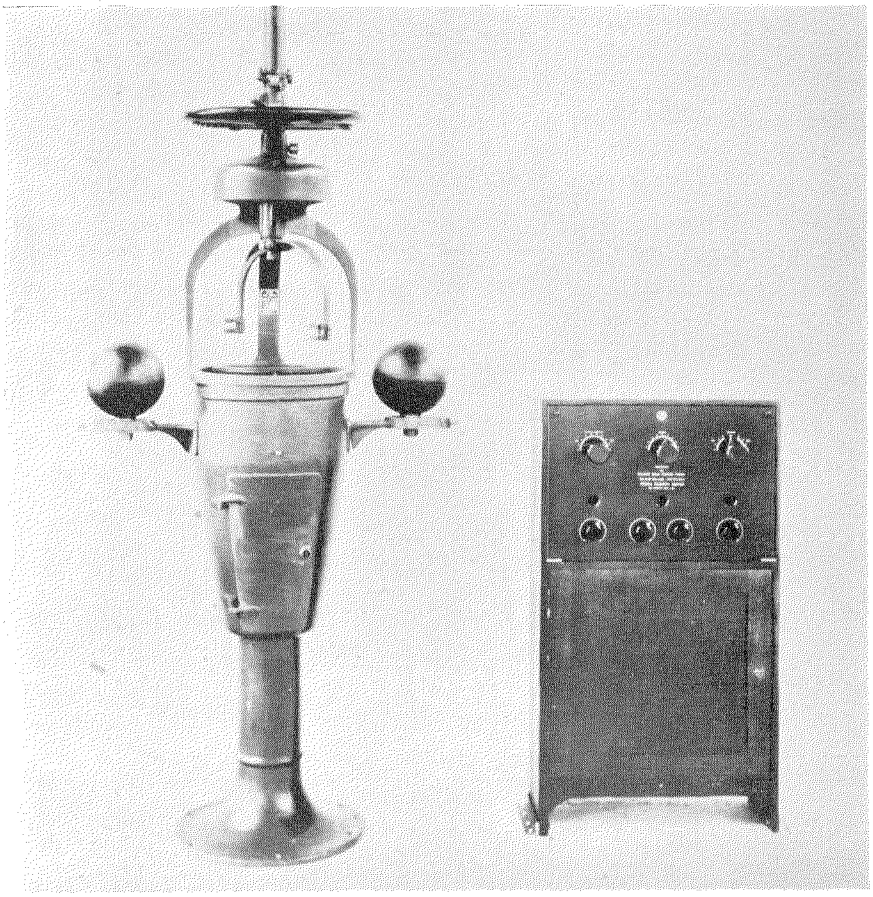
[2] H. Busignies, "The Automatic Radio Compass and Its Applications to Aerial Navigation," *Electrical Communication*, v. 15, pp. 157–172; October, 1936.

[3] "Electrical Communication: 1940–1945," *Electrical Communication*, v. 23, pp. 214–216; June, 1946.

[4] H. Busignies, "Applications of High-Frequency Solid-Dielectric Flexible Lines to Radio Equipment," *Electrical Communication*, v. 22, n. 4, pp. 295–301; 1945.

June, 1921.[5] In 1923, marine direction finders were placed on the market by the Federal Telegraph Company of Palo Alto, California.[6] They consisted, basically, of a rotatable shaft that carried a loop and a pointer moving in front of a magnetic-compass dial. Remarkably enough, despite all progress made in radio since that time, all marine direction finders now on the market in the United States consist basically of hand-operated rotatable loops; the pointer actuated by the loop shaft moves in front of a fixed scale or a scale controlled by a repeater from the gyro compass.[7] These direction finders cover the frequency range of 285–315 kilocycles per second assigned to this service. The frequency range is often extended to 500 kilocycles to permit taking bearings on ships and fixed stations transmitting within this extra band. The bearing of the radio beacon is determined from the signal nulls corresponding to the well-known figure-of-8 directivity pattern of the loop collector. Proper design, including balance of the loop and sensitivity of the receiver, determine the sharpness of the nulls and the quality of the direction finder. Deviations caused by the ship's structures, which may run as high as ±15 degrees, are compensated by a mechanical corrector in the form of a cam adjusted during calibration. The elimination of sense ambiguity is obtained through the transformation of the figure-of-8 diagram into an approximate cardioid diagram by combining the loop signal with that from a small vertical antenna.

The modern marine direction finder of the rotatable-loop type, using either a tuned-radio-frequency or a superheterodyne receiver, is a reliable instrument which, when properly calibrated and maintained, will provide readings with an accuracy of ±1 degree over 70 or 80 percent of the bearings, and 2 or 3 degrees for the rest of the bearings. Ranges up to 100 or 150 miles on medium-power beacons are currently obtained. Recalibration after months of use will not show more than 0.5- to 1-degree



First marine direction finder placed on the market by Federal Telegraph Company for commercial use in 1923. A similar instrument was demonstrated by F. A. Kolster in June, 1921, on board the U. S. Lighthouse Tender *Tulip*.

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[5] F. A. Kolster and F. W. Dunmore, "The Radio Direction Finder and Its Application to Navigation," *Scientific Papers of the Bureau of Standards*, No. 428; January 16, 1922.

[6] F. J. Mann, "Federal Telephone and Radio Corporation," *Electrical Communication*, v. 24, pp. 396–397; December, 1946.

[7] E. H. Price and W. J. Gillule, "Marine Navigation Aids," *Electrical Communication*, v. 22, n. 1, pp. 56–69; 1944.

difference from previous calibrations, if the superstructure of the ship near the loop is not modified.

Abroad, the principle of fixed loops and rotatable goniometers of the Bellini-Tosi type has often been used, as well as the rotatable-loop principle.

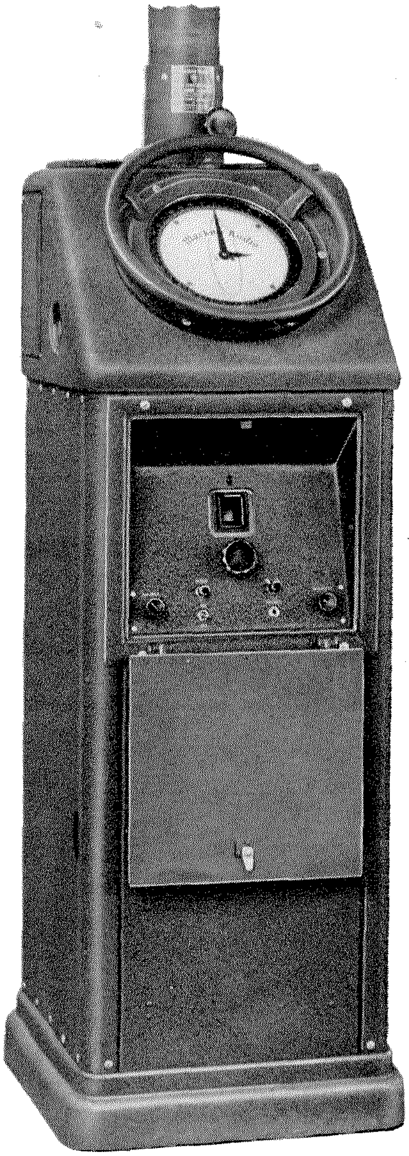
Efforts were made as early as 1926 to produce automatic direction finders requiring only tuning of the receiver to the desired radio beacon for the position of a needle on a scale to indicate the bearing directly. In 1929, a direction finder of the automatic type, developed by the laboratories of the International Telephone and Telegraph System in Paris, was demonstrated on a small private ship in Mediterranean waters. It furnished a direct 360-degree indication without sense ambiguity. A few years later, about 25 such automatic direction finders were installed for normal service on ships of the Portuguese navy. While considerable maintenance was required, the principle of automatic operation without sense ambiguity was completely achieved. Despite these early efforts, no marine direction finder of the automatic type is presently in practical use, so far as current information indicates. This is undoubtedly due to two causes:

A. The apparatus built by the first experimenters was costly and probably too complicated for normal maintenance.

B. The simplicity of operation and reliability of the rotatable-loop direction finder was such that there was very little incentive to change to the automatic type.

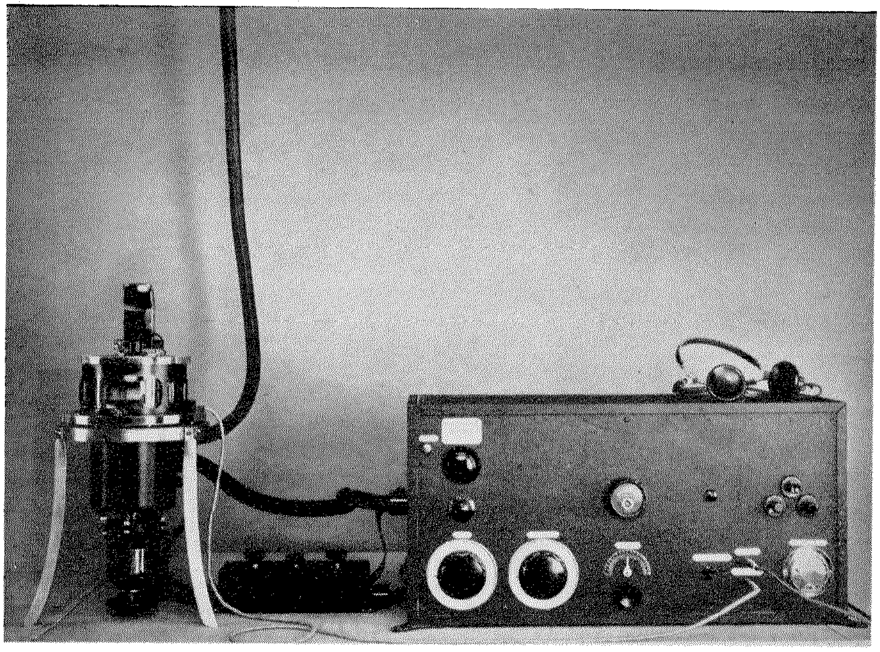
Simplicity is probably also the factor that limited the application of more complex radio beacons, such as those using pulse transmissions, to provide discrimination between direct and ground waves, thus eliminating part or all of the night error that limits the direction-finding range. Considerable research and design work in high- and ultra-high-frequency direction finding was done prior to and during World War II. This resulted in a considerable improvement in the operation of fixed high-frequency direction finders. During the war, the United States Navy installed an important net of such stations, which are now operated by the U. S. Coast Guard.

The Army Air Forces and the Air Transport Command also installed hundreds of instantaneous cathode-ray-indicator direction finders, primarily for guiding and rescuing aircraft, before any other long-range navigational system was available. On the eastern seaboard of the United States, a net of SCR-291 direction finders can be alerted in case of need. Medium-frequency Adock, fixed, direction finders capable of high accuracy at night are also available.



Present type direction finder for merchant marine use.

The night effect is the main limitation on medium-frequency direction finders of the loop type. Direction finders, of course, are not the only radio aids to navigation that suffer from the



First automatic direction finder developed for experimental work in 1928 by the Paris laboratories of the International Telephone and Telegraph System, and tried out on a ship the same year. This direction finder indicated spontaneously on a moving scale the direction of a radio transmission without any ambiguity and throughout 360 degrees. It was the first automatic direction finder developed.

effects of sky waves reflected by the *E* and *F* layers of the ionosphere. Solutions to the problem of night effect that have been applied to shore direction-finding stations have not, so far, been applicable to shipboard installations, because of the size and disposition of the antenna system. Night range is, therefore, limited to something of the order of 50 miles. In the last part of this paper we will return to this problem and discuss possible improvements.

In summary, the first radio aid to navigation, the marine direction finder, is a reliable and simple instrument, still making use of the rotating loop and a means of indication of the same type as was used 20 years ago.

2. *Comparison of Merit and Field of Application with Radar and Loran*

As a result of the inventions and progress made in such pulse-timing systems as radar and loran during World War II, the marine direction finder seems to be subjected to considerable competition. The opinion that the direction finder might become unimportant gained some momentum at the end of 1945 and during 1946, when the first information on the new radio aids was released. At the present time, the situation is much clarified; the performances of radar and loran are well understood. A detailed examination of fields of application shows that marine direction finders, marine radar, and loran are complementary systems, solving different problems of navigation with advantageous overlapping in some cases.

Marine radar is a short-range radio aid to navigation, primarily useful for anticollision service, and for pilotage near the shore or at the entrance to and inside of harbors. A range of 10 to 20 miles (16 to 32 kilometers) is a reasonable figure for its use in these applications.

Loran, on the other hand, is a medium-range (400-mile or 645-kilometer) accurate navigational system, the shipboard operation of which requires somewhat more time and more training than the direction finder. The main question is one of world availability and standardization of loran signals, as operators will have to contend with stations installed in foreign territories in many instances.

Therefore, the marine direction finder, with a day range of the order of 150 miles, is a very excellent intermediate-distance aid to navigation that can bring a ship, with desired accuracy, within the range of operation of marine radar. The advantage of the direction finder for intermediate-distance navigation is simplicity, ease of maintenance, availability, and the relatively small cost of the radio-beacon ground stations. About 190 radio beacons are installed on the shores of the United States and around the Great Lakes. Another 150 are operating throughout the rest of the world. Many countries can increase the services made available at their shorelines at moderate expense.

We shall conclude this section by repeating the more or less accepted opinion that marine direction finders, radar, and loran are complementary systems; that availability of loran transmissions will certainly be the largest factor in determining its generalized use; and that the operation of radar in collecting its information independently of ground or other ship cooperation is the essence of its assured success.

3. *Present Crystallization of Design*

In the first section, showing briefly the evolution and the present state of the art, it was stated that the present marine-direction-finder designs are more or less crystallized around the principle of the hand-controlled rotatable loop, and that the means of measurement by observation of the nulls of the figure-of-8 directive pattern is the same as was used 20 years ago, despite the extraordinary development of radio communication and radio navigation during the same period. The art of direction finding as a whole has progressed during the war to the point where techniques and instruments are available or could be made available in the entire frequency spectrum of radio waves; studies have also been made that greatly improved our knowledge of propagation. Indicating means have been considerably improved; the automatic aircraft compass equips almost every aircraft, the large airplanes even carrying two or three such instruments. Shore direction finders employ cathode-ray tubes for instantaneous presentation of bearings. Ultra-high-frequency direction finders are used in connection with airport traffic control. Moreover, investigation of the applicable patent literature is most revealing; the field of direction finders has not been neglected by inventors. There seems to be no relation between the crystallization of design and the very large number of systems that have been invented. We shall list below the factors that seem to limit introduction of new designs and ideas in marine direction finding:

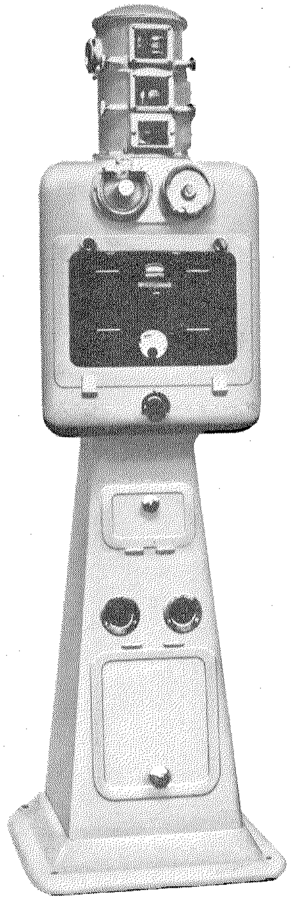
A. Reliability and simplicity of the present design.

B. Availability of a world-wide system of marine radio beacons.

C. Simplicity of the ground beacons and their consequent inexpensiveness.

D. Reluctance and difficulty of introducing changes in the existing world-wide radio-beacon setup.

F. Also, perhaps, the concentration of the attention of laboratories and manufacturers on the problems of radio aids to air navigation.



A commercial version of the completely automatic marine direction finder installed on a number of ships between 1931 and 1934, developed and built by the Paris laboratories of the International Telephone and Telegraph System.

4. *New Approaches by a Definite Departure from Existing Solutions*

Let us next consider whether we have reached a situation where further progress can be obtained at reasonable cost, or whether some new approaches can make the direction finder a still more useful and attractive instrument for navigation.

4.1 FREQUENCY

It might be questioned whether the present frequency range of 290–310 kilocycles is the most advantageous for this service, or could the use of another part of the frequency spectrum result in improved performance with respect to accuracy, range, or reduction of the night effect?

Lower frequencies are eliminated because their inherent characteristics increase the cost of ground stations and the pressing demand for such frequencies by other services; therefore, higher frequencies are suggested. Experience in the field of shipboard high-frequency direction finders leads us to reject that portion of the spectrum for this application. Ultra-high frequencies would definitely be free from night effect, but the range would be limited to line-of-sight conditions; the ultra-high-frequency wave collector would have to be installed on the top of the ship's mast to obtain a 20-mile (32-kilometer) range from a transmitter a little higher than sea level. Despite the increase in accuracy and the elimination of night effect, it seems quite improbable that these would compensate for the increase of cost and, more especially, for the shortening of the range. The incentive for any group to develop an ultra-high-frequency direction-finding system on such a fragile foundation is certainly too weak for a manufacturer to experiment in that direction.

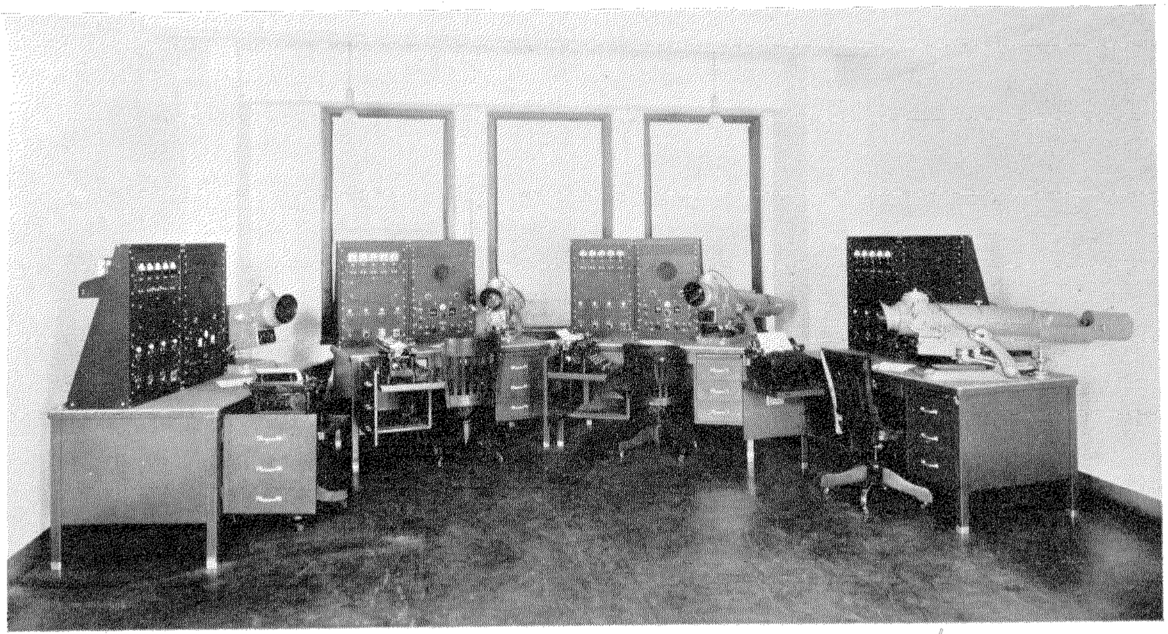
4.2 INSTANTANEOUS DIRECT-READING INDICATOR

Would an instantaneous direct-reading indicator of simple design sufficiently increase efficiency to make it attractive to ship operators and owners, and subsequently justify the effort necessary for its development and adoption?

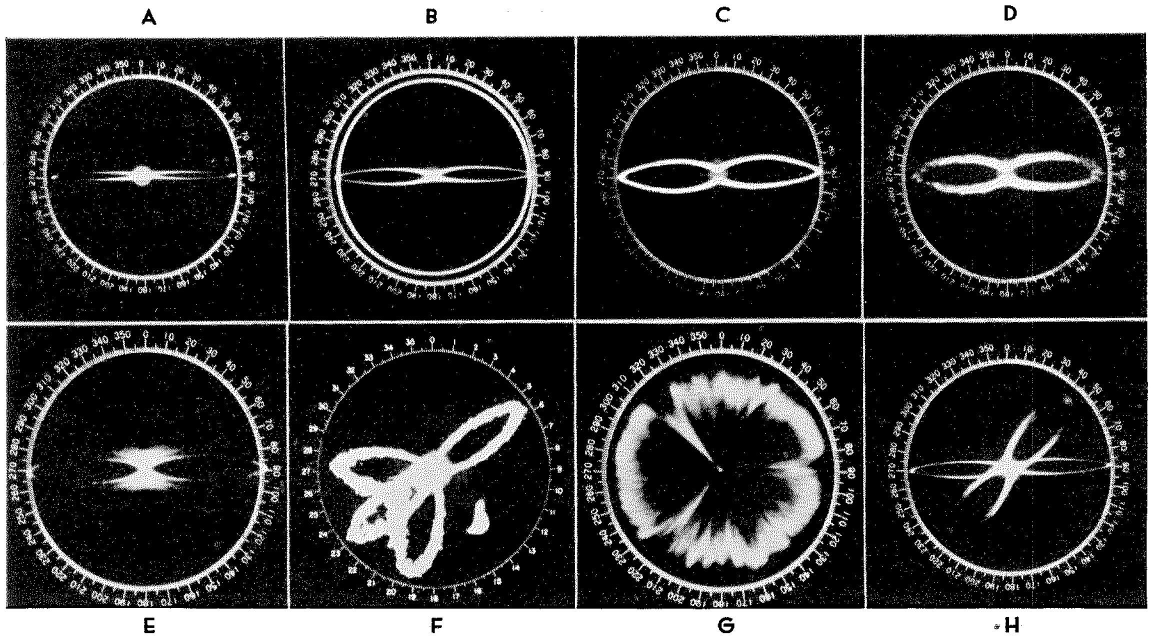
Navigation along the eastern coast of the United States, where beacons are arranged by groups of three on the same frequency and transmit sequentially by units of three minutes, tends to demonstrate that an instantaneous indicator, giving successively the bearings of the three stations, would be quite justifiable on the basis of simplification of procedure. A computer could be designed to facilitate either the recording or the plotting of the fix. The type of indication should be such that a failure would be immediately detectable. It could be a mechanical pointer, a neon or other gas tube, or a cathode-ray tube; however, the simplest and most reliable instrument is to be preferred. In connection with such an indicating system, a null-seeking loop or goniometer, or a small continuously rotating loop or goniometer, is generally required.

While a complete direction finder of such a type has not yet been installed for service tests on board a merchant ship, various components performing equivalent functions have been tested during other developments, and results proved their reliability. This will be an important factor in the choosing among the many possible solutions that offer themselves for direct-reading direction finding. The preferred solutions will be the use of instruments for which a good quantity of data on reliability is already available. Among such we find the null-seeking type of direction finder with mechanical pointer, the rotating loop or goniometer with cathode-ray-tube presentation, and the neon-tube or meter indicators.

Other methods (those avoiding rotating parts) involve electronic goniometers or various



Fixed shore station of high-frequency direction finders for aircraft operated by the U. S. Coast Guard. This direction finder makes use of instantaneous bearing indicators of the cathode-ray type.



Typical patterns of bearing indication shown under various conditions by the cathode-ray direction-finder indicator. *A*—Strong continuous-wave signal. *B*—Strong keyed continuous-wave signal. *C*—Weaker signal. *D*—Weak signal with noise. *E*—Modulated signal. *F*—Superimposed "direction" and "sense" patterns for weak signal with noise. *G*—Noise only. *H*—Two simultaneously keyed signals.

modulation schemes that eliminate the need for three-channel receiving systems.

Generally speaking, whatever the method used to obtain the directive information through one receiving system, the indicator can be of any of the three most important classes just mentioned.

Thus, there is room for ingenuity in design to satisfy the requirement for simplification of procedure.

4.3 FREQUENCY STABILITY AND BANDWIDTH

Could more frequency stability in the transmitter and receiver permit utilizing channeling principles, as in communication systems, or as proposed for other navigational systems?

The resulting simplification of the tuning of the receiver to the radio beacon should rightly be an added feature of a direct-reading direction finder. Stability has already been achieved in many beacons, and the cost of stability in the receiver could be kept reasonably low.

This has another aspect: The bandwidth generally agreed upon in direction-finder receivers at present is many times larger than necessary, even for the purpose of identification of the radio beacon. The audio-frequency modulation of the beacon is not basically required for bearing indications. Provided sufficient stability is assured in the transmitter and in the direction-finder receiver, the present bandwidth of the receiver could be divided by 10; which is equivalent to an increase of 10 times the transmitter power. Much further improvement would still be possible, but might require costlier design. Further advantage can be derived from this increase of sensitivity by increasing the accuracy and stability of the direct-reading direction finder.

Very small bandwidths, of the order of 20 radio-frequency cycles, are being tried in connection with present developments of long-range navigational systems.

4.4 PULSE TRANSMISSION

Could night effect be reduced by the use of pulse transmission?

Pulse transmission for the purpose of discriminating between the ground wave and the sky wave was proposed and tested in Germany prior to the war. Expected results were obtained; the disadvantage being in the bandwidth required for the transmission. The circuits for separating the ground waves from the sky waves in the receiver would definitely be more expensive and difficult to maintain than the simple circuits of the present receivers. Despite its promising aspects, this system would require a considerable amount of further development before a sound opinion could be formed concerning its practical value. It is still doubted that such an improvement, although resulting in increased range at night, would justify itself. However, the question would merit much more detailed consideration if advantage could be taken of the pulse transmission from the ground beacon for the purpose of distance measurement.

Any such modification of the present radiobeacon system on a world-wide basis would meet with difficulties of application that can readily be imagined, and these would contribute to discouraging the inventor or manufacturer.

4.5 DISTANCE-MEASURING SYSTEM

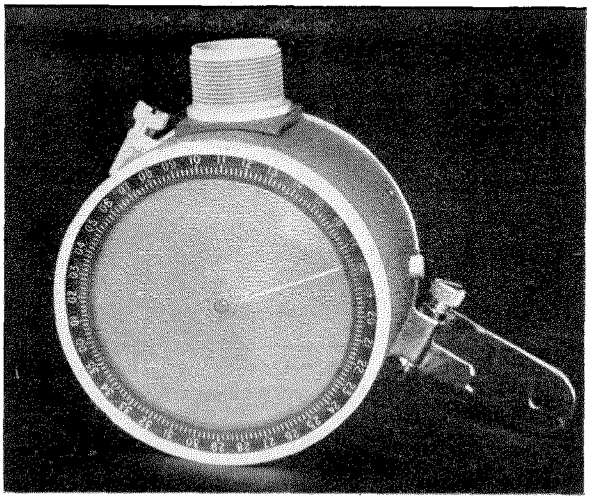
Is the combination of the marine direction finder with some distance-measuring system desirable and/or possible?

In air navigation, the R-Theta system has reached justified success due to many advantages, two very important ones being:

A. A complete fix can be obtained from one station and supposedly from one transmission.

B. The accuracy of a fix is the same at all points in the 360-degree coverage of each station.

Distance-measuring equipment making use of ultra-high frequencies would not be of much assistance to the marine direction finder because its range is limited to line-of-sight. However, one can imagine a medium-frequency distance-measuring system using phase measurement of an audio-frequency modulation for the determination of time of travel, and, therefore, distance. The ship could interrogate the radio beacon on a medium-frequency channel assigned for that purpose; the audio-frequency signal received at the ground beacon would be retransmitted through the beacon to the ship, where the audio-frequency phase-displacement measurement, resulting in distance indication, would be effected at the output of the direction-finder receiver. The drawback to this system is that only one ship at a time could interrogate the beacon; but this interrogation should not need to last more than a number of seconds for each call, and could be limited in time. An indication of the availability of the channel for such interrogation could be provided, with self-identification of transmission. This distance-indicating system could be introduced in the form of attachments that would not require world-wide acceptance before advantages are realized from the first installations. Considerable developmental work would have to take place to solve such a problem properly, and it is very doubtful that a private enterprise would decide to start such a project unassisted.



A straight-line automatic bearing indicator applicable to most types of automatic direction finders.

5. *Conclusions*

Among the possible improvements to the marine direction finder that may be recognized in the not-too-distant future are automatic indication of bearing and a reconsideration of the stability and channeling of radio beacons. Moreover, it seems that the combination of the shipboard direction finder with some form of a distance-measuring system would result in an attractive navigational scheme, upon the development of which some imaginative effort could be applied in the next few years.

H. T. Kohlhaas, Editor, Retires

H. T. Kohlhaas has retired after twenty-three years as editor of *Electrical Communication*. Under his guidance, the magazine became a leading contemporary technical journal of communication development. It is among the leaders in circulation outside the U.S.A. Mr. Kohlhaas' retirement follows more than forty-two years with International Telephone and Telegraph Corporation and its predecessor companies. In addition to his editorship, he was assistant vice president of the Corporation and in charge of its Technical Publications Division.

From 1905 to 1922, Mr. Kohlhaas served in the Western Electric Company Laboratories (later Bell Telephone Laboratories) as an engineer, section head, and finally as a division head in the physical laboratory. In 1922, he was appointed executive personnel assistant to the systems engineer of the Bell Telephone Laboratories. He was transferred to the Western Electric Company headquarters in 1924.

In 1925, Mr. Kohlhaas was assigned to statistical studies and the editing of *Electrical Communication* for the International Western Electric Company. Through the purchase of that company, he became a member of the staff of the International Telephone and Telegraph Corporation. He edited *Electrical Communication* in London from 1937 through 1939 and, while there, served as publicity representative of the International Telephone and Telegraph Corporation and the International Standard Electric Corporation. He was appointed an assistant vice president of International Telephone and Telegraph Corporation in 1945.

Mr. Kohlhaas received a B.S. degree in electrical engineering from Cooper Union in 1907 and later was awarded electrical engineering degrees by that institution and Brooklyn Polytechnic Institute. Mr. Kohlhaas is a Member of the American Institute of Electrical Engineers and a Senior Member of the Institute of Radio Engineers. He is also a member of the Norwegian Chamber of Commerce.

Mr. Kohlhaas has agreed, at least for the present, to continue as consulting editor for *Electrical Communication*. This journal owes a great debt of gratitude to Mr. Kohlhaas and, while we wish him every enjoyment in his well-earned retirement, we also look forward to his continued interest and help as consultant.

FRANK C. PAGE

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S. E. Buckley was born in London on November 28, 1904. He received the B.Sc. degree from London University in 1924.

He joined the European engineering department of International Standard Electric Corporation in January, 1926, as a student engineer, and was transferred to the development and research laboratories at Hendon in 1929. He was again transferred in 1932 to Standard Telephones and Cables, Limited.

Associated with the development of Permalloy-type materials since 1927, he is at present responsible for development and engineering of magnetic materials including dust cores, loading coils, and telephone-cable accessories. Mr. Buckley is in charge of the chemical and metallurgical laboratories at the Woolwich plant.

Henri Busignies

He is an Associate Member of the Institution of Electrical Engineers.

• • •

Henri Busignies was born in Sceaux, France, on December 29, 1905. He received the degree of Electrical Engineer from Paris University in 1926.

On completion of his military service, he entered the Paris Laboratories of the International Telephone and Telegraph Corporation in 1928. Until about 1940, Mr. Busignies traveled to Italy, Spain, Switzerland, England, Africa, and the United States in the interests of the company. Since 1940, he has been an executive engineer and is now the Director of Federal Telecommunication Laboratories in New York city.

His first patent was issued in 1926 and has been followed by almost a hundred others. They have been dominantly in the field of automatic direction finding with particular attention to aircraft applications. He received the Lakhovsky award of the Radio Club of France in 1926.

Mr. Busignies is a Fellow of the Institute of Radio Engineers.

• • •

E. Serra Calabuig was born at Játiva (Valencia), Spain, in 1894. In 1910, he joined the State telegraph services. He continued his studies and obtained the degree of Ingeniero de Telecomunicación in 1920.

He became a plant engineer for the Compañía Telefónica Nacional de España in 1924 and is now director of construction and maintenance.

• • •

J. M. Clara was born at Zaragoza, Spain, in 1892. He entered the State telegraph services in 1908 but continued his studies and received the degrees of Licenciado en Ciencias Químicas in 1914, Licenciado en Ciencias Físicas in 1916, Ingeniero de Telecomunicación in 1920, and in 1923 Ingeniero Radiotelegrafista ESE Paris.

In 1924, he joined the Compañía Telefónica Nacional de España as a district engineer and is now chief engineer. Since 1930, he has represented the Compañía Telefónica Nacional de España on the Comité Consultatif International Téléphonique.

E. Serra Calabuig

• • •

Thomas C. Clark was born in Chicago, Illinois, on January 15, 1921. He received the B.S. degree in electrical engineering from Illinois Institute of Technology in 1942.

On leaving college, he accepted an engineering position with Federal Telephone and Radio Corporation and, from 1945 to 1947, worked on the development of airport traffic-control receivers. In October, 1947, he joined the engineering staff of Air Associates, Incorporated, in Teterboro, New Jersey.

J. M. Clara

THOMAS C. CLARK

Mr. Clark holds an Associate membership in the Institute of Radio Engineers.

• • •

A. G. CLAVIER was born in Cambrai, France, in 1894. He received a degree in electrical engineering from Ecole Supérieure d'Electricité in 1919 and then joined the staff of engineers organized by General Ferrié at the Etablissement Central de la Radiotélégraphie Militaire. He was in charge of research on high frequencies from 1920 to 1925.

In 1929, Mr. Clavier joined Les Laboratoires Standards in Paris which later became Laboratoire Central de Télécommunications, and has been continuously engaged in research on centimeter and millimeter waves. He was in charge of the experiments which, in 1930, resulted in 17-centimeter-wave transmission across the English Channel and of the developments for the Lympne-St. Inglevert microwave radiotelephone link, which was inaugurated commercially in 1934. He was assistant director of research in 1945, when he was transferred to Federal Telecommunication Laboratories in New York, where he now holds the same position.

A. G. CLAVIER

He has published extensively on high-frequency oscillators, wave guides, and general electromagnetic theory, and has taught field theory and applications of ultra-high frequencies at the Ecole Supérieure d'Electricité.

Mr. Clavier is president of the section of the Société des Radioélectriciens dealing with hyperfrequencies. He is a Fellow of the Institute of Radio Engineers, and a Member of the Institution of Electrical Engineers.

• • •

EIGIL COHRT was born on November 19, 1902, in Copenhagen, Denmark. He studied at the Royal Technical College in Copenhagen from 1922 to 1926, when he joined the technical staff of the Bell Telephone Manufacturing Company in Antwerp.

Mr. Cohrt was transferred in 1938 to Standard Electric A/S in Copenhagen, where he became systems engineer and head of the circuit division.

• • •

ROBERT A. FELSENHELD was born on February 15, 1910, at East Orange, New Jersey. He has been active in the broadcast field since 1927. In 1941, he joined the staff of Federal Telecommunication Laboratories, Incorporated, to work on ultra-high-frequency antennas and allied subjects.

Mr. Felsenheld is a Member of the Institute of Radio Engineers.

• • •

SIDNEY FRANKEL was born on October 6, 1910 in New York City. Rensselaer Polytechnic Institute conferred three degrees on him: the B.A. degree in electrical engineering in 1931, and in mathematics the M.A. degree in 1934 and the Ph.D. degree in 1936. He was an instructor in mathematics from 1931 to 1933.

For a year after leaving college, Dr. Frankel served as a sound-recording engineer with the Brooklyn Vitaphone Corporation. In 1937–1938, he was an assistant engineer in the design and development of electronic flight instruments for the Eclipse Aviation Corporation.

EIGIL COHRT

He joined the Federal Telegraph Company staff at Newark, New Jersey, in 1938 as an engineer on the design and development of radio transmitters. In 1943, he was transferred to Federal Telephone and Radio Laboratories, now Federal Telecommunication Laboratories. At present he is engaged in the development of components for microwave systems.

He is a member of Sigma Xi and a Senior Member of the Institute of Radio Engineers.

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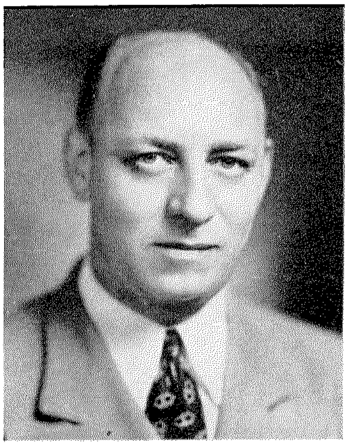
FREDERIK HEEGAARD was born on January 24, 1898, in Copenhagen, Denmark. He received the M.Sc. degree in electrical engineering in 1926 from the

ROBERT A. FELSENHELD



SIDNEY FRANKEL

Royal Technical College in Copenhagen. He then joined the engineering staff of the Post and Telegraph Administration and is now chief engineer of the broadcast service.

He served as vice chairman of the electrical section of the Society of Danish Engineers from 1935 to 1937, and as a member of its board of directors until 1941. He was chairman of the technical committee on Radio Broadcasting House. He is a member of the Academy of Technical Science and of its acoustical laboratory. He has been secretary of the Danish national committee for Union Internationale Radio Scientifique since 1938.

Mr. Heegaard was chief of the Danish delegation to the Bucharest radio conference in 1937, and a delegate to the International Telecommunication Conference in Cairo the following year and in Atlantic City in 1947. He was appointed a member of the Patentboard in 1942.

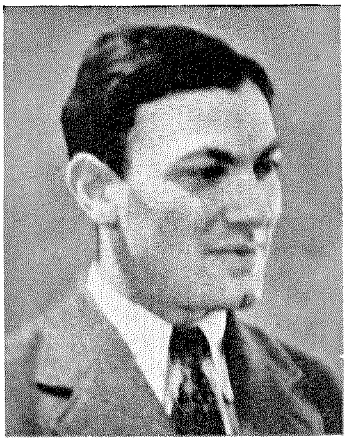
FREDERIK HEEGAARD

Knighthood in the Order of Danneborg was conferred on Mr. Heegaard in 1941.

* * *

HEINZ K. HENISCH was educated at Carlsbad, Czechoslovakia, and Ottershaw College, Chertsey, England. He then studied physics and mathematics at Bristol University and later at the University of Reading, where he obtained the degree of B.Sc. (First Class Gen. Hons.) in 1942.

He joined the radio department staff of the Royal Aircraft Establishment during the war, and is now again at Reading University, engaged on semiconductor research.



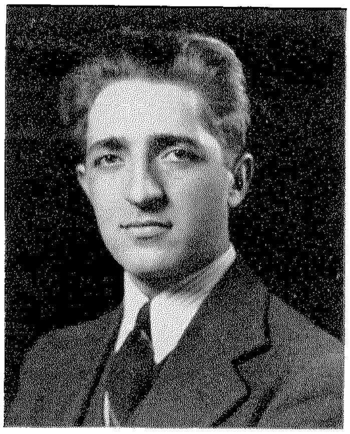
HEINZ K. HENISCH

Mr. Henisch became an Associate of the Institute of Physics in 1944.

* * *

ARMIG G. KANDOIAN was born on November 28, 1911 in Van, Armenia. In 1934, he received the B.S. degree and a year later the M.S. degree in electrical communication from Harvard University.

He joined the engineering staff of International Telephone and Telegraph Corporation in 1935. With its establishment in 1937, he was transferred to the International Telephone Development Company for work on air navigation and instrument landing systems. Federal Telephone and Radio Corporation took over the activities of that company and was succeeded by Federal Telecommunication Laboratories in 1945. Mr. Kandoian is now a department head responsible for radar, antennas, and multiplex broadcasting.



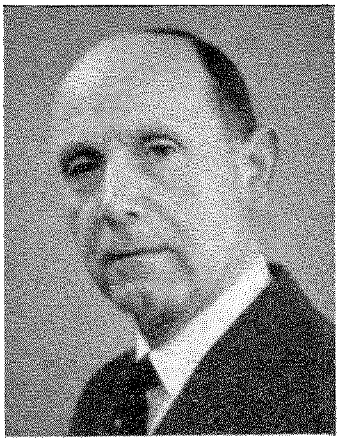
ARMIG G. KANDOIAN

Mr. Kandoian is a Senior Member of the Institute of Radio Engineers, an Associate of the American Institute of Electrical Engineers, and is a member of the Harvard Engineering Society and of Tau Beta Pi.

* * *

G. KLINGELFUSS was born in Aarau, Switzerland on October 19, 1895. After graduation from the technical school in Berthoud as an electrical engineer, he entered the Bell Telephone Manufacturing Company at Zurich in 1920.

In 1921, he was transferred to the engineering department in Antwerp,



G. KLINGELFUSS

Albert Shadowitz

and in 1923 assigned to sales engineering in the branch office at Berne. He served as commercial director in charge of the Swiss business from 1930. In 1941 the Berne branch was attached to Standard Telephone et Radio S.A., Zurich. Since 1935, he has been a member and secretary of the Board of Directors of the Zurich company.

• • •

Albert Shadowitz was born on May 5, 1915 at Brooklyn, N. Y. He received the B.E.E. degree from Polytechnic Institute of Brooklyn in 1935. After attending Harvard University during 1936 and 1937, he studied at the University of California from 1937 to 1940.

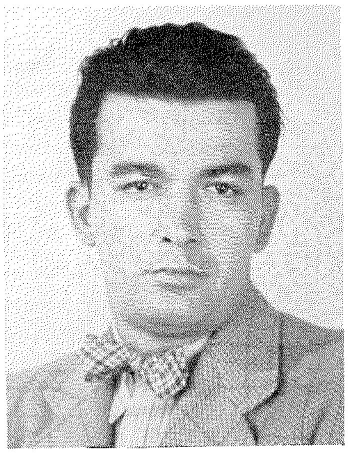
He served as an engineering employee at Aberdeen Proving Grounds from 1941 to 1943. In 1943, he joined Federal Telecommunication Laboratories as an engineer.

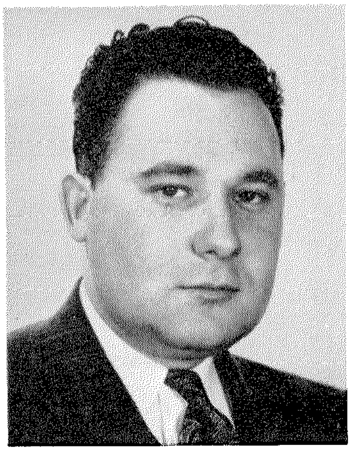
Mr. Shadowitz is an Associate Member of the Institute of Radio Engineers.

• • •

William Sichak was born on January 7, 1916, at Lyndora, Pennsylvania. He received the B.A. degree in Physics from Allegheny College in 1942.

From May, 1942, to November, 1945, he was engaged in developing microwave radar antennas at the Radiation Laboratory, Massachusetts Institute of Technology. Since November, 1945, he has been with Federal Telecommunication Laboratories, Incorporated, working on microwave antennas and allied subjects.


William Sichak


Martin Silver

• • •

Martin Silver graduated from New York University in 1941 with a Bachelor of Engineering degree. He then came to Federal Telecommunication Laboratories where he worked on communication-type jamming equipment. At the termination of the war, he was assigned to the development of frequency-modulation broadcast transmitters, monitors, links, and associated equipment.

Mr. Silver is an Associate Member of the Institute of Radio Engineers.

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Compañía Internacional de Radio Boliviana, La Paz, Bolivia

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Compañía Internacional de Radio, S.A., Santiago, Chile

Radio Corporation of Cuba, Havana, Cuba

Radio Corporation of Porto Rico, San Juan, Puerto Rico[1]

[1] **Radiotelephone and Radio Broadcasting services.**

Cable and Radiotelegraph Operating Companies

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The Commercial Cable Company, New York, New York[2]

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All America Cables and Radio, Inc., New York, New York[4]

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[2] **Cable service.** [3] **International and Marine Radiotelegraph services.**
[4] **Cable and Radiotelegraph services.** [5] **Radiotelegraph service.**

Laboratories

International Telecommunication Laboratories, Inc., New York, New York

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Standard Telecommunication Laboratories Ltd., London, England

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ELECTRICAL COMMUNICATION

Technical Journal of the
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Modern Research Facilities of Federal Telecommunication Laboratories

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WITH the opening of the 300-foot aluminum-sheathed microwave research tower on May 19, 1948, at the new Federal Telecommunication Laboratories in Nutley, New Jersey, a unique approach to high-frequency radio propagation experimentation was inaugurated. The new tower constitutes a research laboratory in the sky; its 5 upper levels and roof extend between 200 and 300 feet above the surrounding terrain. It may, therefore, be described as a building on top of a tower, the top level at a height equivalent to that of a 30-story building.

Like light rays, ultra-high-frequency radio waves travel in substantially straight lines making their range virtually limited by the horizon and they are, therefore, most effectively propagated from and received at places located high in space, particularly for communication over any appreciable distance. Heretofore, sites atop mountains and tall buildings have been employed for microwave research, or the antennas have been placed on tall masts. But mountain tops are inconvenient for day-to-day experiments because of their remoteness and inaccessibility as well as their susceptibility to the extremes of weather. Building tops also present many limitations, such as interference caused by the reflection of waves off nearby structures, possible disturbance of existing radio services, and the difficulty of complying with local governmental regulations. Masts, because of their instability, inaccessibility of the structures mounted on top of them, need for supporting wires, and the necessity for long transmission lines between apparatus and antennas, are poor devices for precise measurements.

Tower in Original Plans

When plans for the new laboratories were first drawn up in 1943, a rigid tower for microwave experiments was included. At that time, the laboratories were busily engaged on radar, high-frequency direction finding, and other similar wartime developments. Since then, interest in microwave techniques has increased rapidly, not only for radio navigation with such devices as distance-measuring equipment, assisted radar, and Navar, but also for television, especially ultra-high-frequency color television, frequency and pulse-type modulations, and facsimile systems for broadcast, point-to-point, and relay operation. An extensive microwave program has been adopted by the armed forces of the U.S.A. and during the next decade will involve contracts with private industry of about a thousand million dollars for research and development. It is evident, therefore, that the concept of a tower as an essential unit of a complete electronics laboratory was both correct and foresighted.

The laboratories as first conceived were to consist of two major divisions, a central part supplying a number of common services and a group of research units. In the common section were to be included technical and administration, maintenance, security and personnel, main incoming power, heating, fire detection and protection, compressed air and gas, and similar services. The first idea was to house these common services in a single building as a center surrounded by the laboratory and research buildings. The tower was to be located on top of the central building.

At this point, architects were consulted and the problem was presented as one of engineering a highly functional group of structures combining extreme flexibility and low maintenance. The first design drawn up by the architect provided for a circular central building three stories high with laboratory wings on either side. The tower was to be an open steel structure with two glass-enclosed floors at the top. This first plan would have permitted a total of seven laboratory wings

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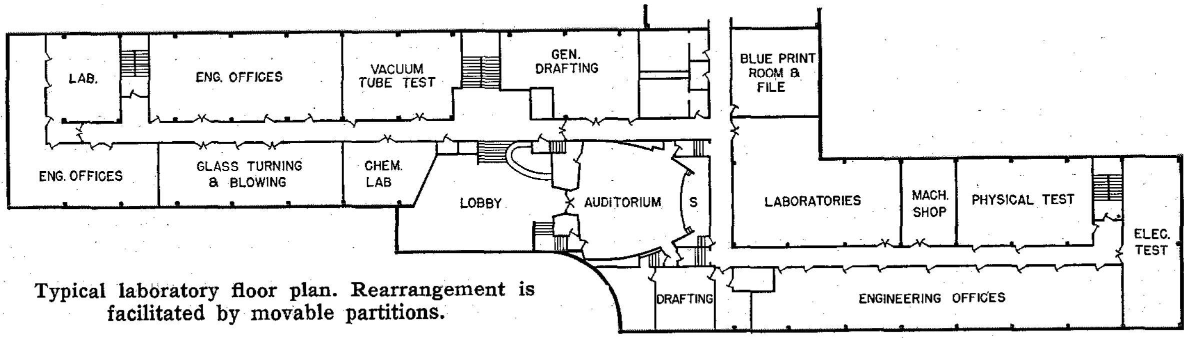
New buildings and tower of Federal Telecommunication Laboratories are shown on the opposite page.

to be built radially at 45-degree angles to each other; the eighth radial would have been omitted to provide for the main entrance.

When this initial plan was presented, it became apparent that it had a number of disadvantages. First, as wings were added, the windows on the first floor of the central structure would have been blocked out so that ultimately the whole floor would have been without outside lighting. Secondly, construction of the steel tower on top of the main building would have had to proceed at the time of the erection of the main building. The tower could not have been built afterwards because of the danger and difficulty involved in carrying on construction while the building below was occupied. Thirdly, the appearance of the open structure on top of the building was considered unsatisfactory. Fourthly, the architect recommended against wings at 45-degrees from each other because such construction introduced complications in the steel framework.



Typical laboratory floor plan. Rearrangement is facilitated by movable partitions.

Functional Design

In considering the original design further, it was felt also that the general architectural conception was not modern enough and that, as long as new laboratories were being constructed, they should be as functional as possible and should indicate some degree of independence from the classical method, provided there was justification for such a departure.

After several modifications and revisions were considered, a plan for the buildings was approved. The plan provided for an oblong central building with a group of laboratory wings at either end. In these later plans, provision was made for an independent tower to be erected close enough to the central building for easy access, but far enough away to be constructed independently. However, since the permanent tower could not be built immediately owing to governmental wartime building restrictions, no further designs were considered at the time. Work on the first two buildings was started in November, 1944, following the approved plan. These buildings were occupied the following spring. To provide immediate facilities for microwave research, a 200-foot temporary wooden tower was built nearby.

When the first two buildings neared completion, it was decided to plan for two additional wings. As this construction was to proceed after V-J day, it was possible to include a number of items that could not be considered during the war, that is, an appropriate entrance lobby, an auditorium, and the permanent tower. The four buildings now completed provide 130,000 square feet of gross floor space and comprise the largest group of metal-sheathed buildings in the world. Flexibility has been provided in the building site plan so that additional buildings may be added without impairing the symmetry of the group.

Aluminum for Outer Walls

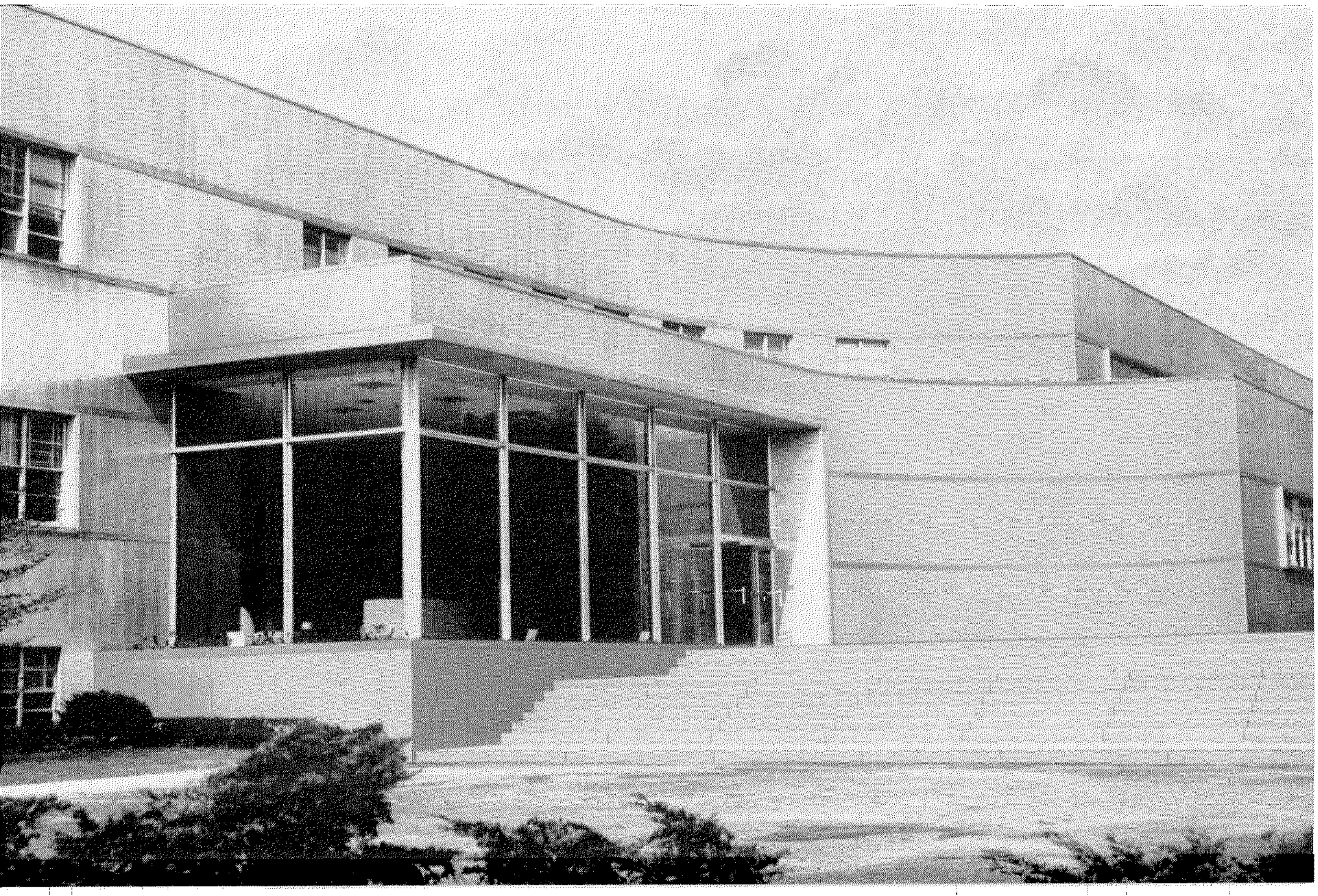
Although utilizing conventional steel framework, the buildings depart in other ways from normal constructional design. The final plans specified a large amount of glass for the laboratory exterior plus a suitable material in the form of plates for the outside walls. Among the materials considered were an asbestos-cement compound and enamelled steel.

It was found however that at the particular period of the war when the buildings were to be started, an excess of aluminum plates was available. Investigation of aluminum for the outside wall covering revealed that samples of this material submitted to exposure in an industrial area from 10 to 15 years tarnished progressively, but not to objectionable proportions. After this period of time, it is possible to restore the original appearance by application of a suitable finish. Actually, such refinishing may not be necessary since the action of wind and dust on the buildings seems to constitute a continuous polishing agent. Also, it was realized that the aluminum surface would give a fair degree of shielding to the inside rooms against external radio fields produced by nearby radio stations, which measurements showed to be of the order of one-tenth volt per meter. Shielding between laboratories was also effected through the use of metallic partitions. With the use of all-steel flooring of the Robertson type, shielding between floors would also be effected.

The exterior walls, therefore, consist of panels made up of an outer covering or skin of aluminum, an insulation space filled with Fiberglass, and a backing of steel. These panels are two feet wide and are erected vertically, usually as spandrel sections.

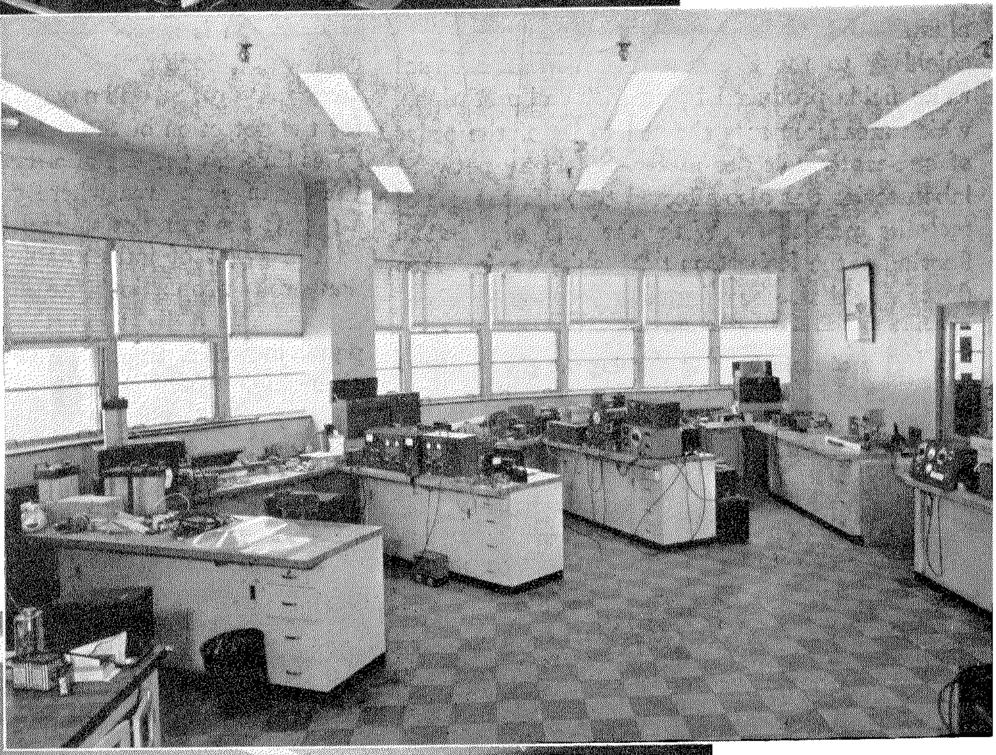
The ceilings of the buildings are of the suspended type, utilizing metallic acoustic panels, which may be removed whenever access is desired to the ceiling space. The space between the suspended ceiling and the underside of the floor immediately above is utilized for air ducts, special ventilating equipment, sprinkler pipes, and other service and heating pipes.

Provision is made for distribution of hot and cold water, compressed air, illuminating gas, nitrogen, hydrogen, and oxygen. There are facilities for the detection of the presence of inflammable gases through the use of a detector that samples the air in each area where gasses are used and analyzes the samples several times each minute. In the event of inflammable gases escaping in dangerous proportions, an alarm automatically sounds and a light on the main gas-detector panel indicates the space in which the leak is occurring.



Front view of main building showing the glass-enclosed entrance lobby.

The auditorium may be used for lectures and demonstrations or as a studio for sound and television broadcasting.



Typical laboratory showing the large amount of natural illumination provided.

Typical office area. The wall panels in the rear can be moved to expand or contract the space as may be required.

Flexibility with Six-Foot Module

To obtain maximum flexibility, a system of removable dividing partitions is used. These allow the size of rooms to be changed with relatively little effort and without destroying the wall material. A six-foot space module was established. The first building, constructed during the war, employed Transite wall paneling for interior wall liners, while the later buildings have steel wall liners. The module was respected in designing heating, lighting, and sprinkler systems so that each module is self-sufficient in these respects. In other words, modules can be combined in any desired arrangement without the need of changes in these three basic services.

A maximum of natural lighting is provided through use of large window areas. Artificial lighting, in addition to conforming to the six-foot module, provides complete flexibility within a unit. The basic lighting unit is a fluorescent fixture with two four-foot tubes set flush with the metal-pan acoustic ceiling. The fixtures are of the snap-in type and utilize the same supports that carry the hung ceiling. It is a simple matter to replace two ceiling pan units with a lighting fixture at any location.

Varying from usual practice, the light switches are located in the corridor. At each column, there is a two-foot wall section that is not movable and provides a fixed position for these switches. Thus, partitions may be moved without involving electrical changes. The switches are, however, coordinated with the six-foot module system.

Electrical distribution throughout the laboratories is by means of a four-wire grounded-neutral system supplying 208-volt, three-phase 60-cycle alternating current. Thus, 120-volt, single-phase alternating current may be obtained between any phase wire and neutral. A substation is provided for each of the buildings and for the tower. The primary voltage of each substation is 4150 volts. Facilities are available for supplementary substations should the load require additional power capacity.

Fire Prevention System

The sprinkler system was the subject of special attention, since there are few places where water would cause more loss in both time and money than in an electrical laboratory. The rate-of-rise maximum-temperature system was, therefore, installed to prevent unnecessary water damage. In this system, the rate of temperature rise actuates the controls rather than some predetermined fixed temperature. Any abnormal temperature rise would start the following chain of protective measures.

A. Fire is detected by rate-of-rise heat detectors.

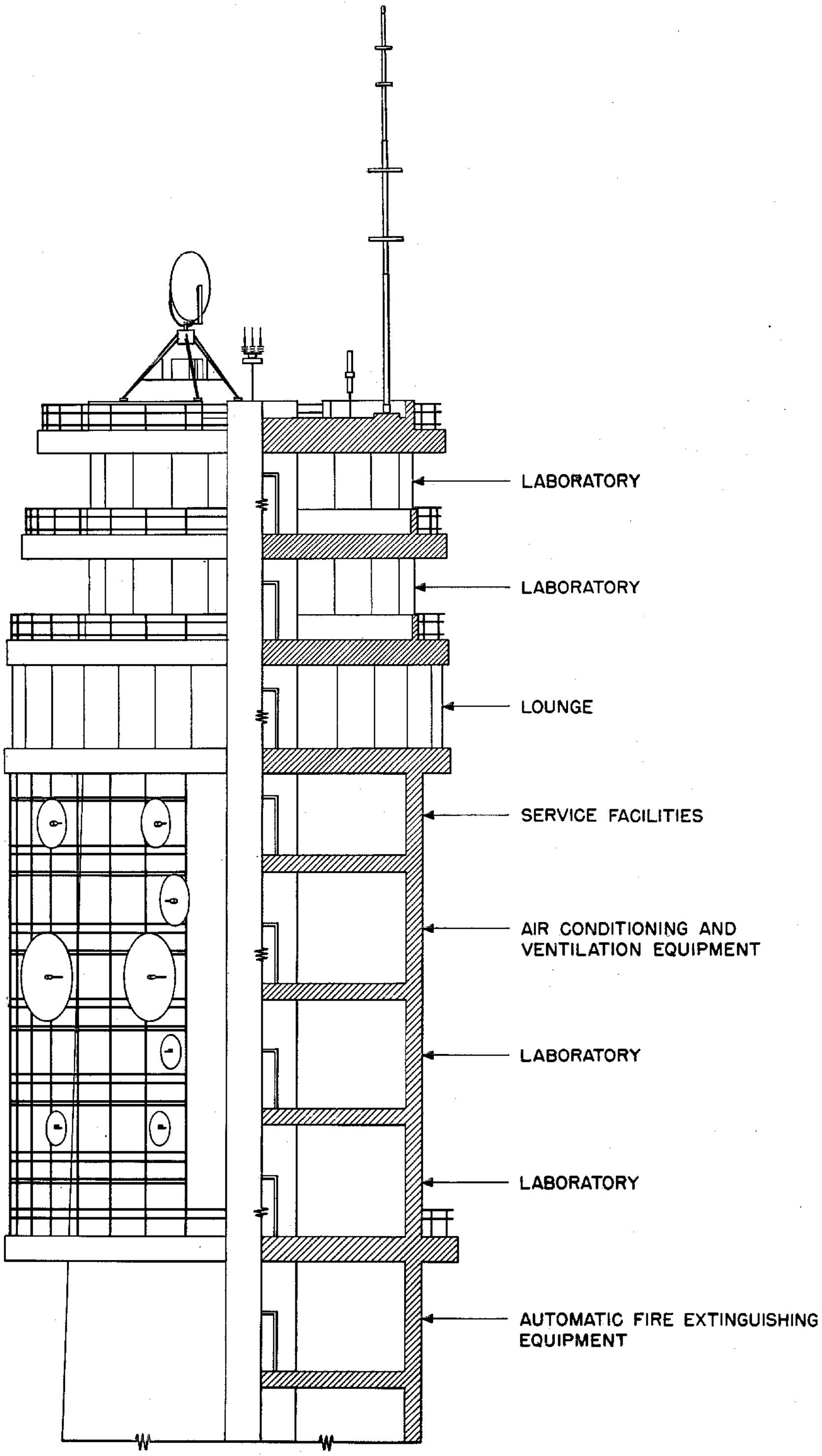
B. An alarm is sounded, indicating the fire zone in which the trouble is located.

C. Water is allowed to enter the sprinkler lines.

D. The water is held back until either
    a. manual extinguishing efforts succeed, in which case the sprinkler heads do not open and water damage is prevented, or
    b. the sprinkler links fuse and water under high pressure is sprayed in the affected area.

The heating plant is a steam and high-vacuum system using fuel oil. Controls permit the adjustment of the temperature in each room.

Corridor of one of the laboratory buildings.


LABORATORY
LABORATORY
LOUNGE
SERVICE FACILITIES
AIR CONDITIONING AND VENTILATION EQUIPMENT
LABORATORY
LABORATORY
AUTOMATIC FIRE EXTINGUISHING EQUIPMENT

Detail and cross section of upper levels of tower.

At the time the lobby and auditorium were being planned, two new basic designs for the tower were also drawn up. One was for a square main shaft with square buttresses and the other was for an oval structure. The towers in both cases were to be totally enclosed providing ample working space in a series of levels. Both designs appeared to have merits, but the oval shape was chosen because it lent itself more readily to experiments in all directions while the square shape might have made it difficult or impossible to operate in certain directions.

Optimum height of the tower for the location was determined through experimentation on the 200-foot structure. These experiments showed that heights between 200 and 300 feet would yield satisfactory results for research in all but one direction. This height would not be great enough for signals to scale the Watchung Mountains to the west. A tower tall enough to do this was considered economically and structurally impractical. Since research was to be carried on chiefly in the southeasterly direction toward the International Telephone and Telegraph Building in lower Manhattan (New York City) and the field station at Telegraph Hill (Hazlett), New Jersey, the 300-foot height was chosen.

Tower Design Unique

The task of designing a high steel tower with a comparatively narrow base presented a number of interesting problems. Foremost among these was that of weight distribution. With 75 percent of the weight concentrated between the 200- and 300-foot levels, the necessity to provide adequate support was a major consideration. For example, architectural engineers estimated that during high winds the uplift stress on the foundation might well run into several hundred tons. As a result of these calculations, a tension anchorage was developed that is probably the most unusual feature of the entire construction.

The main columns of the tower are anchored in a reinforced concrete mat 10 feet thick. Uplift loads are carried by reinforced tension cylinders extending approximately 30 feet below the foundation mat into bedrock. The rock layer in this area is said by geologists to be thousands of feet thick.

To construct the anchorage for the tower, a well driller drove 24 holes each 14 inches in diameter into the ground. Twelve large steel rods were lowered into each hole and a special type of concrete, which expands on hardening, was used to anchor the cylinders in the surrounding rock.

The frame of the tower, including columns and bracing, is made of structural steel of standard rolled shapes. Roof and floor slabs are reinforced concrete on steel beams; the upper levels employ slightly different framing because of depressed floors and overhanging balconies.

The south balcony at the lower landing is supported on a cantilever extending 12 feet beyond the column supports and the resulting uplift on the main floor girder is reacted by an inverted K-frame to the floor above. The south columns above the main landing, having a design loading of over 200 tons, are carried back to the main columns by cantilever girders. Resulting uplift on the main floor girder is counteracted by a tension tie to a K-frame two stories below.

Rigidity of Tower Essential

Because of the exacting nature of the work to be conducted in the elevated laboratory—notably the use of sharply focused beams—rigidity of the supporting structure was of paramount consideration. A primary design requirement, therefore, was limitation of both static deflection and amplitude vibration under a wind velocity of 80 miles per hour. This was dictated primarily to avoid any major interference with focussing operations during stormy periods.

Accordingly, the allowable wind stress in the columns and bracing was so calculated that, with the maximum wind load from any direction, total deflection at the top of the tower would not exceed 10 inches. These calculations, taking into account the unusual weight distribution at the top of the tower and the eccentric loaded projection on the north-south axis, involve both normal and torsional inertia loads as well as actual wind load.

In considering these special conditions, original formulas had to be developed to approximate the amplitude of vibration that would result from periodic and unbalanced wind pressures that would add to the static deflection of the tower under a steady wind load. The period of oscillation of the tower involved calculations based on inertia forces of the elastic center of gravity and the elastic center of oscillation. This was computed to be about 3 seconds, exclusive of the damping effect of the bracing system and the applied wind load. On the basis of these assumptions, it was determined that the maximum static deflection would be about seven inches in addition to an amplitude of vibration of about three inches.

For beam antennas, vertical supports are provided between the lower landing and the main landing, standardized as to spacing and serviced by steel walkways. The pipe supports are attached to the cantilever balcony curb. At present, they are installed only in the southeast quadrant, but provisions have been made for setting them up all around the tower. A cargo boom located on the roof permits a parabola or other type of antenna structure to be hoisted to the point of location. The accessible walkways then permit easy location of the antenna in any chosen position. The boom is also used to haul materials to the roof.

An automatic electric elevator with a speed of 500 feet per minute provides rapid access to the upper levels. The car has a capacity of nine persons or 2500 pounds and will accommodate standard 29-inch bays for transportation to the research levels. Each research floor provides 1000 square feet of laboratory space.

Practical as well as aesthetic considerations dictated the choice of aluminum as the outside covering for the tower. Rolled sheet aluminum sections not only provide the required enclosure for elevator and stairwell, but afford weather protection to the steel frame, thus reducing maintenance to a minimum.

The laboratory space is steam heated on the two upper landings, while the roof has been provided with snow-melting pipes in the concrete. Both electricity and hot and cold water are supplied to these areas. Extensive fire-protection equipment has been provided throughout the tower. The water pump, elevator relay controls and motor generator, flood-light controls, substation, and other service equipment are located in a lower part of the tower. Water is pumped to

Former club house of the Yountakah Country Club now used as a restaurant and recreational center by laboratory personnel.

a pressure distribution tank located at the 230-foot level and distributed from there.

Glass-Enclosed Landing

The main landing is entirely glass enclosed and air conditioned. This area forms an observation room providing a wide view of New Jersey, New York harbor, and the New York City skyscrapers. Decorated in a modern style, in keeping with the impressive functional design of the tower and buildings, this level may be used to demonstrate equipment as well as for research. Included in the decorative scheme is a large mural, symbolizing the progress of electronic science through the centuries, painted on stainless steel by Mrs. Buell Mullen, well-known designer and artist who introduced the technique of painting on stainless steel. A number of her murals are on exhibition in the Library of Congress and in other governmental buildings in the U.S.A. and also Latin America.

A floodlight trench circles the tower at its base to provide projected incandescent illumination, while a blue cold-cathode light strip extends vertically to the top. The tower, then, becomes an impressive beacon at night, its spectacular beauty dramatically mirrored in a large, gracefully curved reflecting pool at its northern edge.

The laboratory buildings and tower are located on the site of the former Yountakah Country Club, whose rolling lawn and tall, stately shade trees have been preserved. The lawn provides auxiliary outdoor areas for field tests and experiments with mobile and portable equipment.

Located several hundred feet to the southeast of the laboratories is the club house of the former country club. A large rambling colonial-style building, it now houses the restaurant and cafeteria for the laboratory employees. Also, there are spacious rooms, which are used for employee recreation. Directly adjoining the club house is an outdoor swimming pool available to laboratory personnel for evening swims during the summer months. An outdoor dance pavilion may be used when picnics are held in the grove. Behind the club house are areas for softball and other games.

Electronic Carillon

A 25-note electronic carillon has its loudspeakers attached to the topmost railing of the tower. With an audible range of approximately 12 miles, its musical tones may be heard in the surrounding community as well as on the laboratory grounds. At hourly intervals the carillon tolls Westminster chimes and musical selections are played on it each noon and evening. So, besides being an efficient facility for research, the tower fulfills the function of the campanile, familiar to old-world communities and now present in some form on almost every college and university campus in the U.S.A.

Less than 20 miles from New York City, surrounded by pleasant and healthful countryside, and in the midst of desirable home communities, the new Federal Telecommunication Laboratories are located in an atmosphere stimulating to scientific development. Since the products of a laboratory are discovery, invention, and design—creations of the intellect—the environment for its scientists and engineers is as important as the tools with which they work. In the new laboratories, the two needs may be said to have been ideally fulfilled, for not only has an advanced, functional, and well-equipped group of buildings been provided, but they have been placed in a setting that makes for a pleasant place to work and live.

Intelex — Automatic Reservations

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Standard Telephones and Cables, Limited, London, England

COMPARISON of Intelex with existing manual methods of reserving passenger space shows outstanding improvements. While the manual systems may require even hours to confirm a reservation, Intelex will make and confirm a reservation within 10 seconds, and since substantially all operations are performed by circuits using well-tried relays, switches, and tubes, the probability of error is greatly reduced. A brief coded message of reservation, cancellation, or inquiry may be transmitted by teleprinter from any local office. No further manual operations are required and the Intelex machine will reply automatically to the teleprinter in the office originating the order. Information can be obtained at any time on the status of all reservations on any specific journey.

• • •

The growth of transportation in recent years has been accompanied by such a vast multiplication of the complexities of the services, that manual methods of handling many of these are becoming inadequate. A service seriously affected by this phenomenon is the reservation of seats for passengers on airlines and railroads. Diligent efforts by the managements, as well as expenditure of ever-increasing amounts of money, have not enabled the carriers to keep pace with the mushrooming permutations and combinations of space demands brought about by the increasing needs of the traveling public.

Although all forms of transport, including railroads, airlines, busses, and steamships are feeling the pinch of this problem, it is, perhaps, most critical in the airlines. Cost of operation of aircraft is so great that filling of every possible seat on every flight is imperative. At the same time, to sell more seats than are available is disastrous to public relations. The handling of 15,000,000 passengers per year, filling every possible seat but not overselling, is therefore a difficult manual operation.

To accomplish this task, one major airline employs about 700 persons in its reservation service at a cost of about a million and a half dollars per year—nearly 3 percent of the total cost of that airline's operation. Nearly half a million more is spent by this airline for the communications facilities needed to handle these reservations.

To help reduce this huge cost and at the same time improve the reservation service to the passenger, a new system of handling reservations, called Intelex, has been developed. By replacing much of the personnel by electronic and electromechanical devices, it provides great advantages in speed and accuracy, reduces costs, and improves customer relations.

Intelex is essentially an electromechanical "brain" arranged to provide a bookkeeping service, the basic requirement of a reservation system. In its "books," or "ledgers," reservations and cancellations are recorded from many remote points. Information is also provided regarding various totals of the entries. Intelex will record and confirm any reservation or cancellation within 10 seconds. Through the "Have-a-Look" feature, the status of seats on any trip may be determined also within 10 seconds.

In the following paragraphs, a comparison is made of the facilities available in reservation systems as used commonly today, with the features of Intelex. Special attention is given to airlines, which must meet particularly difficult requirements.

1. *Reservation Systems in Common Use*

1.1 GENERAL

An airline system carrying passengers in either direction between two stops once a day (Fig. 1) has a relatively simple job when organizing a reservation system. There are then only two journeys a day, one each way. Space must be provided in the reservation record, or ledger, to enter the sale of, say, 30 seats per day on each of two journeys, and up to perhaps 8 days in advance. The number of seats in the ledger is then 2 × 8 × 30, or 480.

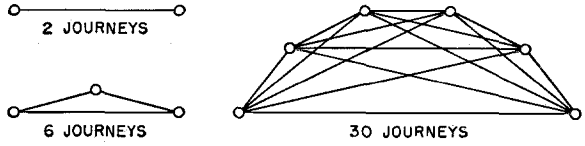


Fig. 1—Number of journeys possible in transportation systems having 2, 3, and 6 stops. Each line joining two stops indicates two journeys, one in each direction.

If a third stop is added to the route (Fig. 1), the number of possible journeys (between any two stops) per day advances from 2 to 6, and the ledger entries for the 8-day advance booking become 6 × 8 × 30, or 1440 seats. Similarly, if the number of stops is doubled from 3 to 6, 7200 seats are listed for 8 days. It is, therefore, apparent that the complexity of the ledger entries increases tremendously with the number of stops on a particular route; the number of journeys possible in both directions on a route having n stops is $n(n - 1)$, indicating that the journeys increase almost as the square of the number of stops.

It will also be apparent that the entries increase directly with the number of flights made each day over the same route, with the number of different routes, with the number of seats on each airplane, and with the number of days of advance booking allowed.

Several domestic airlines of the U.S.A. are each carrying over 1,000,000 passengers a year, and some considerably more. A total of 15,000,000 passengers are carried yearly.

Each of these sales must be recorded, and the number of such entries is complicated by the fact that many reservations are cancelled and the space resold before flight time. Rough estimates, allowing for the fact that during certain days and certain hours the number of transactions are at a peak, show that the average for each 100 scheduled flights with approximately 100,000 passengers per month, involve 1000 messages in the busiest hour. This means that receiving and distributing the messages in a manual centralized office, finding the correct ledgers, and making the notations must be carried out at a speed of one message in every 4 seconds.

Furthermore, the many transactions described must be made from a large number of places. A typical large airline may have a hundred or more district or local reservation bureaus, airport offices, and ticket offices from which reservations are made.

Intelex equipment racks in typical arrangement for a small airline.

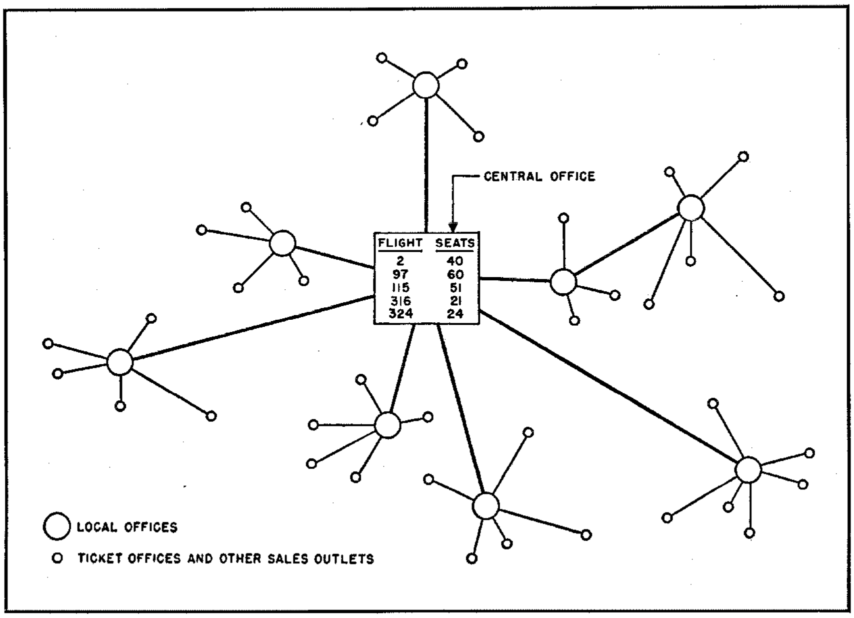


Fig. 2—Typical centralized reservation system. Each local office requests seats from the central office, where the records are kept.

Even with 700 persons handling reservations, it has been found that as long as two hours are often required between the request for a reservation and the confirmation that space is available.

1.2 Request-and-Confirm System

Reservation systems may be classified by the method of handling requests, such as "request and confirm" or "sell and record." Under the first system, the request of the customer is relayed to the headquarters where the ledgers are kept, a search of the ledgers is made, and a message is sent back to the ticket office stating either that the reservation has been made, or that the space is not available. This method, were it not for the time factor and the possibility of errors through the human element, is the most accurate; no space is sold until its availability has been verified. When the traffic becomes heavy, as it is today with nearly 15,000,000 air passengers per year, there is simply not enough time to search the ledgers for each reservation, enter the item, and return a confirmation or denial message. The problem of time is not serious when reservations are requested a week or so in advance, but it becomes more and more acute as flight-time is approached; while this last is to some extent true of any method, it is particularly true here.

Thus, although the request-and-confirm system was the first to be used, and is the most accurate, it has generally been replaced by the sell-and-record method.

1.3 Sell-and-Record Method

Developed primarily to speed up the service to customers, sell-and-record is based on the use of a space-availability board, or chart, in each local reservation office. By reference to this board, clerks can give immediate answers to most of their customers' requests.

Under this system, now in use by most major airlines, local offices inform headquarters each time a reservation is made. Entries are made in ledgers at the headquarters, and the local offices are informed when the flights are almost filled up in order to keep the availability boards up to date. Time is saved since no return confirmation or denial message is used. As flights approach capacity, the facilities again become jammed, since the local office must then, because of possible delay in posting the availability board, call headquarters to see if sufficient space is left, and an answering message is returned. Although a great deal of work has been saved in the long run, mistakes are more likely to occur than with the request-and-confirm method. Selling of the same space twice may occur, or unsold space may be left. Failure to receive a "sold" message in the central office may go unnoticed until it causes oversale. All these difficulties are particularly likely to happen during the last few hours before flight time.

1.4 Centralized Versus Decentralized Systems

Practical reservation systems may be classified as above into sell-and-record, request-and-confirm, or any combination of those two methods, but another useful classification depends on the location of space-control activities. Systems are either "centralized" or "decentralized" under this classification, depending on whether all of the space on a single flight is tabulated and controlled at a single point or is distributed among several points.

1.4.1 Centralized

A diagram of a typical centralized system is given in Fig. 2. A central reservation office, probably near the traffic center of the network, contains the ledgers in which the reservation entries are made for the entire system. This office is connected by teleprinter or telephone lines to each local office in the system. In areas where traffic is relatively light, several local offices may be connected into a single telegraph circuit. As an indication of the number of personnel and the space required, a major airline may use about 200 persons in the central office, and about 25 in each of 20 local offices.

Fig. 3 shows the flow of information in a typical centralized system. In each local reservation office are teleprinter machines connecting to the central office, a large space-availability board, and a number of incoming local telephone lines, some connected to various ticket offices, and some with numbers listed in the local telephone directory for direct calls from customers.

Reservations made by the local offices are sent into the central office until the flights are almost full, perhaps 3 or 5 seats below capacity. The central office then "broadcasts" this fact over a second teleprinter line to each local office. When the flight is actually full, this is also broadcast. On the flight-availability board mentioned above, colored tags are systematically arranged on hooks to indicate the "condition" of the flight, i.e., whether reservations can be confirmed immediately, space has to be requested from the central office, or whether no space can be sold.

To trace the path of a reservation, let us assume that a customer telephones the local office requesting a seat on an airplane traveling between stops *A* and *B* three days hence, leaving at 3:00 P.M.

The clerk who receives this call in the local office will look at the tag that corresponds to this particular date and flight on the availability board (Fig. 3). If the tag indicates that space is available, he will inform the customer that he may have the seat, and asks him when he wishes to pick up the ticket. The information relative to his reservation is noted and placed on file. By means of the teleprinter, the central office is notified of the sale of this seat. In the central office, clerks enter the information in the partcicular ledger corresponding to the flight and date.

When the customer arrives at a ticket office to pick up his reservation, the ticket clerk telephones to the local reservation office and gives the flight, date, name, and terminals (*A* and *B*). When the proper record is found, i.e., the card that was filed above, the information is checked, the card is marked "Ticketed" and placed on file, and the ticket clerk sells the ticket.

When the customer phones in for a reservation, if the clerk in the local office should find from the

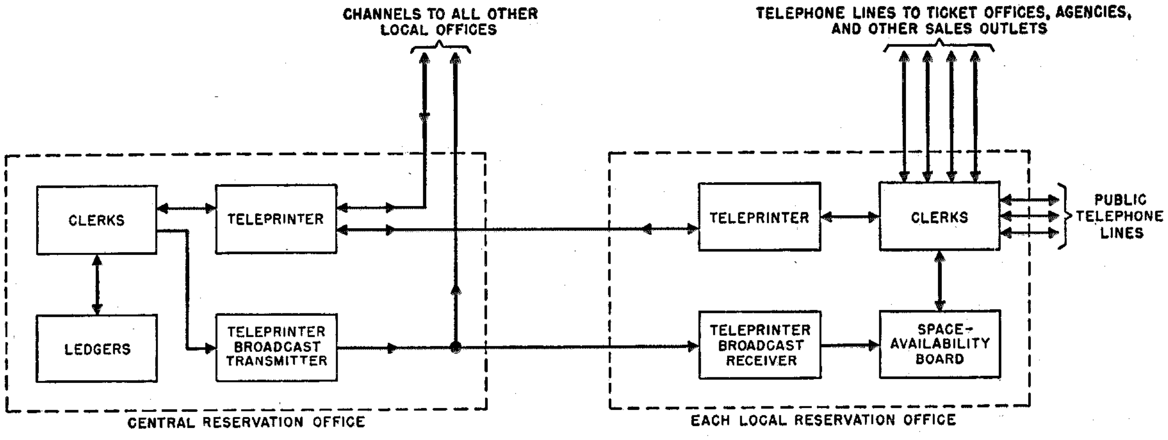

CHANNELS TO ALL OTHER LOCAL OFFICES
TELEPHONE LINES TO TICKET OFFICES, AGENCIES, AND OTHER SALES OUTLETS
CLERKS
TELEPRINTER
LEDGERS
TELEPRINTER BROADCAST TRANSMITTER
TELEPRINTER
CLERKS
PUBLIC TELEPHONE LINES
TELEPRINTER BROADCAST RECEIVER
SPACE-AVAILABILITY BOARD
CENTRAL RESERVATION OFFICE
EACH LOCAL RESERVATION OFFICE

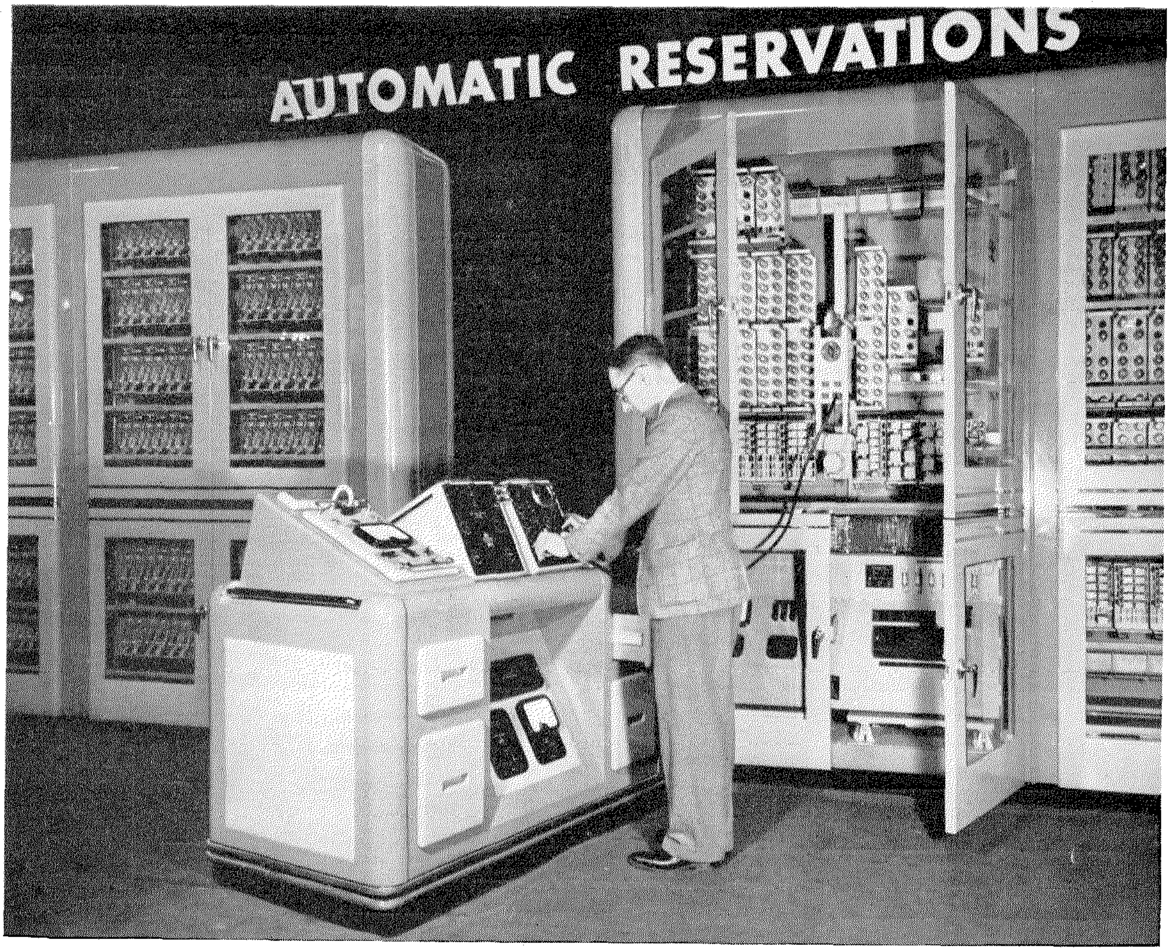
Fig. 3—Reservation network using centralized sell-and-record method, showing flow of information.

availability board that space is not immediately available, he will know that he must check with the central office whether there are any seats left on that flight. The clerk will therefore tell the customer that he will have to confirm the reservation and will call back. After checking by teleprinter with the central office, if the answer states that there is a seat available, the clerk gets in touch with the customer and the procedure is as above. It may be noted here that this follows the request-and-confirm procedure mentioned above.

Should the local-office clerk find a "stop-sale" indication on the availability board, or should the above inquiry to the central office show that there are no seats available for the requested flight-time and date, the clerk will offer the customer alternative flights. That is to say, if there are no seats left on the 3:00 P.M. plane, he will offer seats on the next suitable flight showing on the availability board as "available" or "on request."

All of this checking takes considerable time, and it often occurs that cancellations are not sent into the local office in time for resale of the seats. Such time lags immediately preceding flight departure prevent agents from positive selling, and result in space being available which cannot be sold. Therefore, one may often get a seat by waiting at the airport until departure time. This is a common practice, but is unfortunate from every point of view, since the would-be passenger may be disappointed or, if he does not come to the airport, the unsold space reduces flight revenue.

Again, since a customer's reservation may not



Engineer checking Intelex operation by means of equipment on test-set console. The racks in back of the console include switches, called the ledgers, for the storage of reservation information.

be noted in the ledgers for two hours or so after being approved by a local-office clerk, it is apparent that duplicate sales may occur before the availability board can be brought up to date. This is due to the time-lag occuring before the central-office clerk receives that particular notification (he may have many others to take care of first), finds the ledger, makes the notation, and gives instructions for the flight-condition change to be broadcast.

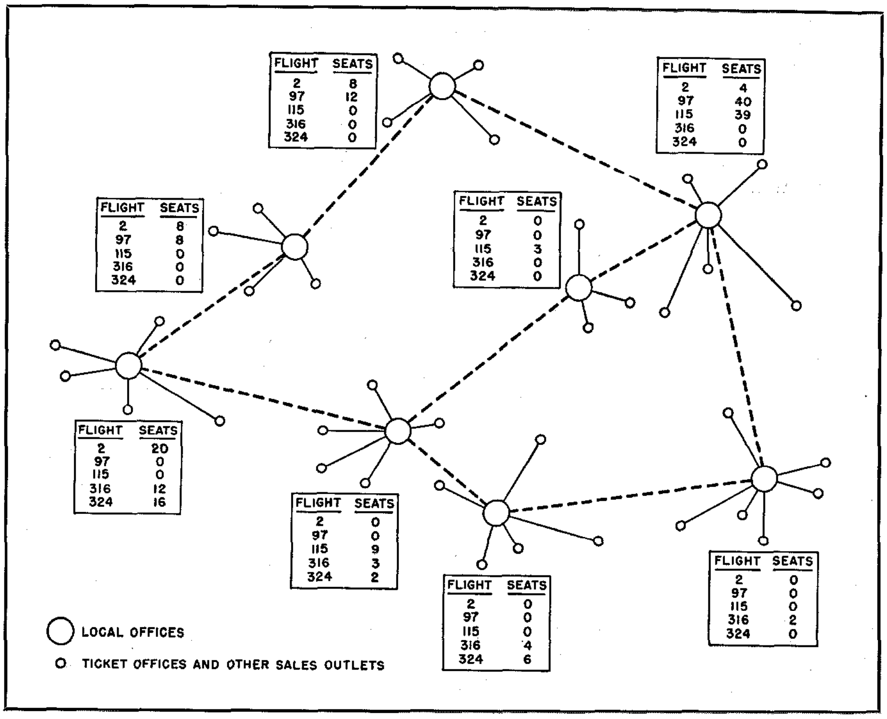


Fig. 4—System having same geographical layout as Fig. 2, but arranged as a decentralized system. Typical allocations of seats among the various offices are shown. Odd-numbered digits travel east to west, and even-numbered go west to east. The dashed lines indicate a communication network necessary for the exchange of reservations among offices having unsold seats and those requiring an excess over their allocation.

It must also be remembered that different offices in any other locality may be giving reservations on the same flight between any of the later stops. Thus it is possible that a passenger wishing to fly from New York to San Francisco may be prevented from doing so by previous sale of all space on any one of the intermediate legs of the flight; such as on the Cleveland–Chicago leg of a New York–San Francisco flight. If the fact of such sale is not known in the New York office, it makes possible, again, a duplicate sale.

1.4.2 Decentralized

Some airlines and nearly all railroads use a method, somewhat similar to the above, that might be called a decentralized, or an allocation system (Fig. 4). Here, the facilities of the central reservation office are divided among the various local offices, the ledgers of the reservations pertinent to the adjacent area being kept in each local office.

Each local office would be authorized to sell a certain number of seats on each trip, based on the average needs of that particular area. The ledgers accounting for the sale of this number of allocated seats are then kept by the local office. The selling procedure is the same as above, except that it is not necessary to send messages to a central office.

However should a local office sell all of the allocated seats, it becomes necessary to get in touch with other local offices to see if they have any to spare. This "shopping" entails considerable long-distance communication between local offices, whereas, with the centralized system, all local offices communicate only with the central office.

If a local office has unsold reservations just before flight time, they are given to a particular office, generally the one at the starting terminal of the flight, for sale there.

Among the basic faults of this system are the duplication of the ledgers in the local offices, since several offices will be handling different numbers of tickets on the same flight (Fig. 4). This may lead either to duplicate sales or to unsold seats. Another fault, mentioned previously, is the great amount of communication between offices in searching for the last few unsold reservations.

2. *Intelex*

The use of Intelex provides a reservation system in which the human element has been reduced to a minimum. A typical system using Intelex is outlined in Fig. 5. This is a *centralized* reservation network that can be operated by the most efficient combinations of request-and-confirm and sell-and-record methods. Adequate protection is designed into this system to prevent most of the difficulties inherent in the manual methods described above. It will be noted, by comparison of Figs. 3 and 5, that the Intelex equipment serves to replace the clerks and ledgers of the central office and most of the teleprinter operators in the central office. In a typical manual centralized system, as many as 30 teleprinters may be required in the central office. From Fig. 5, it will be noted that the operation of the Intelex equipment is performed directly from the teleprinter in the local reservation office. Any printers in the central office are used for monitoring Intelex reservations and for handling "in-the-clear" (uncoded) messages pertaining to reservations or passenger service.

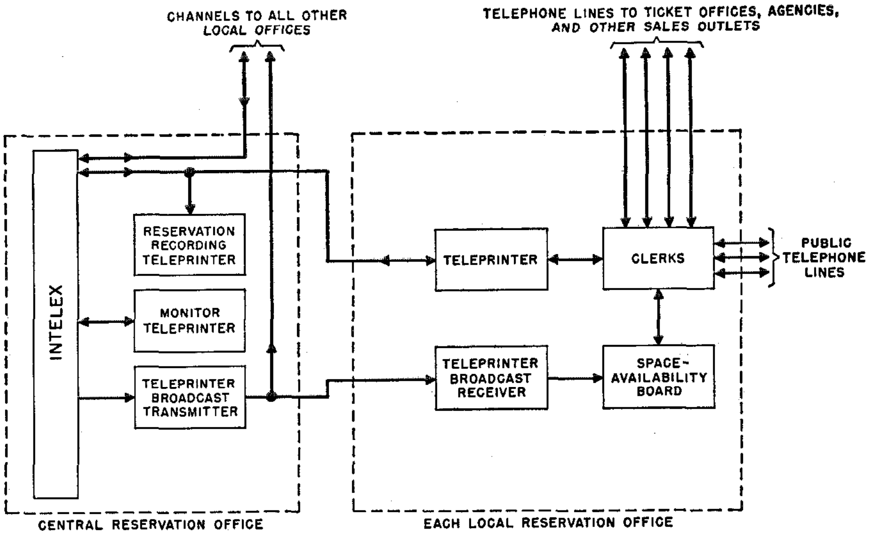

CHANNELS TO ALL OTHER LOCAL OFFICES
TELEPHONE LINES TO TICKET OFFICES, AGENCIES, AND OTHER SALES OUTLETS
INTELEX
RESERVATION RECORDING TELEPRINTER
MONITOR TELEPRINTER
TELEPRINTER BROADCAST TRANSMITTER
TELEPRINTER
CLERKS
PUBLIC TELEPHONE LINES
TELEPRINTER BROADCAST RECEIVER
SPACE-AVAILABILITY BOARD
CENTRAL RESERVATION OFFICE
EACH LOCAL RESERVATION OFFICE

Fig. 5—System of Fig. 3, showing replacement of clerks and ledgers in the central office by Intelex.

The replacement of personnel by automatic equipment (of the type found in automatic telephone exchanges) provides an ultimate in speed and accuracy. It is especially to be noted that when the Intelex system operates on the request-and-confirm basis that was mentioned earlier, it provides the additional advantage that, if no seats are available on the requested flight, the equipment will search other trips to find the best substitute for what was requested and will "offer" this to the requesting office.

The Intelex requires but 10 seconds or less between the start of the requesting message and the end of the answer from the equipment; no human relaying or translation is required between the time when the order message is typed on the keyboard of the local-office teleprinter and when the answer is received on the same machine. A confirmation can be given to a customer within a minute or two from the time that he makes his request.

To handle messages at a speed of one every 4 seconds, or faster as required in the central office of a very large airline, the Intelex equipment is multiplied to provide for the simultaneous handling of a sufficient number of messages to meet the demands of the peak message traffic with a maximum delay of only a few seconds.

2.1 Intelex Central Office

In an Intelex central office, Fig. 5, all of the tabulations of reservations are kept by ledger switches in the equipment (see accompanying illustrations). The additional apparatus in Intelex comprises that used for the choice of the proper ledger switch, and for advancing the switches as each reservation is recorded.

The query from the local-office teleprinter is automatically fed directly into the equipment, and the Intelex answer is automatically sent back to the same teleprinter. Every message to and answer from the Intelex is also printed on a monitor teleprinter that is located in the central office. This teleprinter has two functions: When a query from one of the local offices is unanswerable, being either incorrectly coded or referring to nonexistent flights or impossible dates, the Intelex answer is an *M*, and a signal calls the attention of the monitoring clerk. Also, should an uncoded message be sent, this will not be accepted by the Intelex, but will appear on the teleprinter for the attention of the clerk, who will either give an answer or route the message to the proper office.

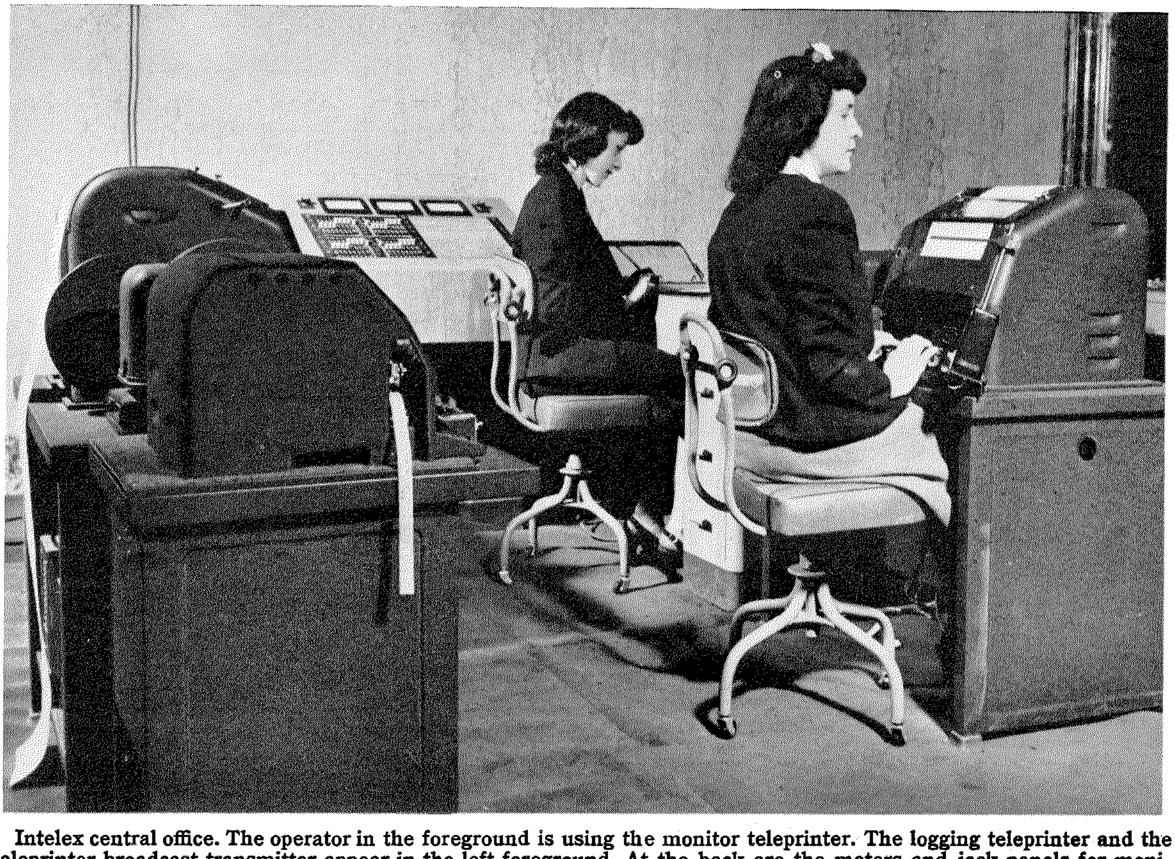
In the central office, there is a reservation-recording teleprinter that prints only the Intelex answers in which a reservation was granted or cancelled.

Also in the central office is the teleprinter broadcast transmitter, which automatically originates and transmits the Intelex information necessary to bring the local-office availability board for any flight into the on-request or stop-sale condition; that is, within, say, 3 seats of capacity, or where no sets are left, respectively. This information, when received in the local office, is manually applied to the space-availability board.

2.2 INTELEX LOCAL OFFICE

In the reservation offices of an Intelex network, the arrangements are substantially the same as those for the centralized manual system described previously. Only a small change in the type of code used is necessary. Because of the speed with which Intelex will answer any request, the reservation will be confirmed practically at once and while the customer is still on the telephone.

The availability board, which was a prime requisite in the case of the pure sell-and-record system, loses some of its importance here, since Intelex can operate on the request-and-confirm basis. However, in some cases it will be advantageous to retain the board as a means of finding the number of an appropriate flight



Intelex central office. The operator in the foreground is using the monitor teleprinter. The logging teleprinter and the teleprinter broadcast transmitter appear in the left foreground. At the back are the meters and jack panels for monitoring and adjusting the operation of the Intelex equipment.

Typical Intelex local reservation office. The clerks seated around the desk have access to the incoming telephone lines. In the center of the desk is an operator at the teleprinter connecting to the Intelex equipment in the central office. The availability board is in the background, with the teleprinter broadcast receiver (not visible) at its right.

with all possible speed. That is, the availability board here replaces a time table or schedule. By retaining the colored tags on the board, a certain percentage of needless messages will not be sent to the machine, and some time will be saved. The information posted on the board, if it is used, is obtained from the teleprinter broadcast circuit.

2.3 An Intelex Reservation

For the purpose of illustrating the operation of an Intelex reservation system, let us again assume that a customer telephones a local office and requests a seat on a 3:00 P.M. plane traveling three days hence between stops *A* and *B*.

On receipt of the message in the local office, the clerk takes the above information and the customer's name and telephone number. The clerk then looks at the board, and if there is an indication of space available corresponding to the requested flight, he confirms the space to the customer, and obtains such other information as where and when he wishes to pick up the ticket, whether he will take the airline limousine to the airport, etc.

While this is being done, the clerk at the telephone gives a teleprinter operator the following information on a small card:

Clerk's identification number.
Flight number.
Day of the trip.
Journey (start and finish airports).
Number of seats.

The teleprinter operator then punches a tape and sends a message, which may be of the following form, to the Intelex central:

C101 F059 D0518 J12 S01 A3 Z

The first group (*C101*) is the clerk's identification number; the second is the flight number; the third is the date (5th month, 18th day); the fourth is the journey designation (corresponding to airports *A* and *B*); the fifth indicates the one seat requested; the sixth, *A3*, informs the machine that the action to be taken is that of a reservation (as against inquiry, cancellation, or other actions); and the *Z* at the last is the termination signal. The starting signal does not appear as a printed symbol. The above message may be prefaced by the customer's name, which the Intelex will ignore, but which will appear on central-office printers as a record for future reference. Also, the exact form of the message may be modified to meet the requirements of a particular airline.

In about 10 seconds, the Intelex equipment will return an answer as follows:

```
N1328
```

The letter and four digits comprise a serial number for the transaction, as assigned automatically by Intelex.

In most cases, even though the clerk has already informed the passenger that his request has been confirmed (on evidence of the availability board), the confirming serial number will be received before the customer hangs up. In case the board had shown the flight to be "on request," the clerk would not inform the customer of confirmation until the serial number was received. This serial number is entered on the reservation card where it becomes proof of a valid, recorded reservation. The card bearing all the information relative to the reservation is then placed on file in the local office until a clerk in one of the ticket offices telephones that the customer has arrived to pick up his ticket. After checking the customer's information against that on the card, the customer will be sold his ticket and the card will be marked "Ticketed" and placed on file.

If no space is available to fill a request on a particular flight, Intelex will search for another suitable flight having a sufficient number of seats available, and will *not* make a reservation, but will return an answer of the following type:

```
F063 D0518 L16/13/15
```

Note that the Intelex has not assigned a serial number to this transaction, serial numbers being given only when a reservation or cancellation is recorded. By the above answer, the equipment informs the local-office clerk that there are 16, 13, and 15 seats left on the various legs of flight 063 leaving on May 18. If the alternative appears satisfactory to the customer, another message will be sent requesting a seat.

Should the board indicate that space is unavailable on the requested flight, the clerk will suggest to the customer such other flights as may be suitable before interrogating Intelex.

2.4 CANCELLATION

One of the great advantages of this automatic equipment is that the ledgers are always up to date. Both reservations and cancellations are immediately reflected in the ledger settings. If a customer cancels his trip through a local office, the clerk there will send a message to the central office, as follows:

```
C107 F153 D0518 J13 S02 A2 Z
```

This message has the same significance as a reservation message, except for the *A2*, which indicates Action 2, or cancel. When the cancellation is recorded in the ledgers, the machine will return an answer as follows:

```
N8678
```

As in the case of a reservation, a serial number has been assigned. Although the above flight has several sectors, the cancellation referred to only the flight sectors between airports 1 and 3. The machine enters this information and changes the availability only on the legs concerned.

2.5 INQUIRY OR "HAVE-A-LOOK" FEATURE

It is also possible to send a message to the Intelex for the purpose of determining the number of seats left on any particular flight. This message would be as follows:

```
F153 D0516 A1 Z
```

No clerk's number is necessary, since there is to be no reservation or cancellation action by the machine, and hence no office record is required. The *A1* indicates that the message is an inquiry.

The Intelex answer might be

```
F153 D0516 L07/06/03
```

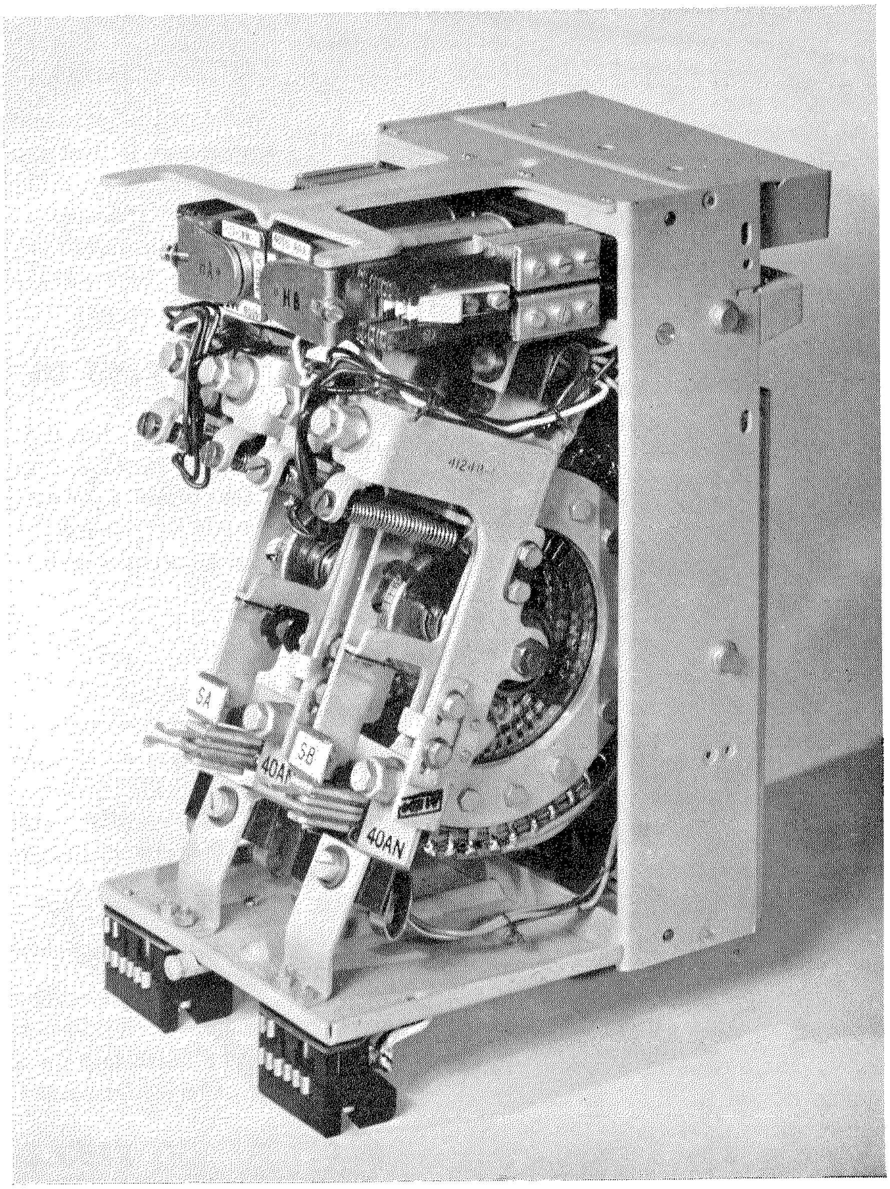
Again no serial number has been assigned. The answer indicates that there are 7, 6, and 3 seats left on the various legs of the flight.

Inquiry messages may be addressed to the machine from any of the offices, local or central. This function is of use for determining the number of passengers on a flight for the purposes of arranging transportation to the terminals, serving meals enroute, determining the weight load, and also where it is desired to see if there is room on a flight for a large party traveling to a single destination. In the last case, it is desirable to determine the seats available on each of several consecutive flights to see if the availability fits the requirements of the party before making the reservations.

2.6 Future Reservations

Intelex, when arranged to handle space reserved up to 8 days in advance, provides storage for 90 percent of all reservations, since only about 10 percent are requested more than 8 days in advance.

When space is requested on flights in advance of this 8 days, the clerk will immediately accept the reservation, since seats are certain to be available on days beyond the storage date. Such reservations can either be held in the local offices until the date of flight falls within the storage time and then passed to the ledgers by normal Intelex methods, or they can be transmitted to the central office at the time of reservation. In the



Ledger switch as used in Intelex. Two switches are mounted on a single jack-in chassis. An advantage of Intelex is that the reservation information may be stored on the contacts of such switches as this. Each switch stores the information for one flight on a given day between two airports. All components in Intelex are mounted on removable chassis for ease in maintenance and for rapid replacement with spares.

latter case, the message will be the same as a normal reservation message (Section 2.3 above), except that the action symbol will be *A4* (future). The Intelex equipment will take no direct action on this message, but will answer with *M*, and operate a signal in the central office. This attracts the attention of the monitoring clerk to the message, which appears on a monitor teleprinter.

The clerk will enter the information in a "Futures" book. When the central-office clerk sets up the Intelex to handle reservations having the requested date of this message, the information in the reservation will be recorded in the machine.

2.7 WAITING LIST

In the same manner as the above, a waiting list may be set up in the central office for customers who wish to be informed if cancellations occur on certain flights. The action symbol used in this case is *A5*. The signal again attracts the attention of the central-office clerk.

When cancellations are made on flights already completely filled, Intelex reports them to the waiting-list agent. By consulting the compiled waiting list for the system, this agent can authorize use of this space in accordance with the policies of the airline.

2.8 RECORDING FACILITIES

Attention may be drawn here to the recording facilities available with the Intelex system. The following services are supplied:

A. Printed records in each ticket office of all messages originating in that office and the answers thereto.

B. Printed records in the central office of all messages transmitted to Intelex and the answers thereto.

C. A selective printed record of all reservations and cancellations on a separate logging teleprinter in the central office.

D. When the reservations on a given flight have brought the total to within, say, 3 seats of capacity, this information is transmitted simultaneously to all ticket offices over the broadcast circuit, and a record of this message is also made in the central office.

3. Conclusions

Intelex is an entirely new means of providing reservation facilities. It is applicable to any system where a limited number of seats are available, and which may be requested, cancelled, or inquired about from any number of different places. It approaches the subject from a modern view using electronic and electromechanical devices. These components have been proven as to stability and long life over a substantial period of time.

The Intelex system is especially noteworthy for its speed and accuracy. These advantages particularly stand out when comparison is made with the manual systems now used by airlines and railroads. Intelex, therefore, permits an efficiency hitherto impossible and hence assists toward the economy of operation that is so important in the transportation business.

Although Intelex is new, the component parts have been tested over the years in automatic telephone exchanges located throughout the world. Many of the operating circuits and features have also been taken from that field, which is noted for its reliability. By concentrating all the equipment in one place, an adequate maintenance program can be followed to prevent failures and interruptions in the operation of the system.

The Intelex equipment is extremely adaptable, since the circuits may be connected to handle any number of flights, for any number of days in advance, and for any number of different airports. The number of possible requesting sources, or local offices, is likewise unlimited. Intelex normally operates from the usual 5-unit start-stop printing-telegraph code.

It may be pointed out here, in conclusion, that the Intelex system, besides its application in industries requiring reservation facilities, such as railroads, airlines, hotels, theatres, and others, has additional possibilities: Intelex is basically a bookkeeping system capable of handling an unlimited number of entries into a great number of different ledgers. Possible additional applications are in the fields of inventory control, credit authorization, chain-store bookkeeping, etc. Intelex would be useful in any system requiring remote-control bookkeeping facilities.

International Telecommunication Convention, Atlantic City, 1947

By P. E. ERIKSON

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THE document that has governed international telecommunication during the last 15 years by agreement between the nations of the world is known as the Madrid Convention of 1932. In July, 1947, plenipotentiaries of 78 nations met at Atlantic City (New Jersey, U.S.A.) for the purpose of revising and bringing up to date that Convention. The new (Atlantic City) Convention is in four parts.

A. Convention proper.

B. General regulations, attached to the Convention.

C. Protocols and resolutions.

D. Agreement with the United Nations Organisation.

These notes are not intended to give a detailed account of the new Convention and its associated documents, but rather to point out important changes or additions to the Madrid Convention.

* * *

1. Convention Proper

The International Telecommunication Union, which will be referred to as the Union, has heretofore had its headquarters at Berne, Switzerland. It has been decided that the Union shall in the future have its seat at Geneva, Switzerland, and the move is to take place by the end of 1948. The former name "Bureau of the Union" will be changed to "General Secretariat."

A rather important article, defining the purposes of the Union, has been introduced into the new Convention. These purposes are:

A. *To maintain and extend international co-operation for the improvement and rational use of telecommunication of all kinds.*

B. *To promote the development of technical facilities and their most efficient operation with a view to improving the efficiency of telecommunication services, increasing their usefulness and making them, so far as possible, generally acceptable to the public.*

C. *To harmonise the actions of the nations in the attainment of these common ends.*

In commenting on the more important changes in the Madrid Convention, the serial numbers of the articles dealt with below are those of the Madrid Convention. The numbering of the articles in the Atlantic City Convention does not correspond to that of the old Convention.

The Madrid Convention (in Article 16) provides for the setting up of international consultative committees and the annexed regulations recognise as existing three such committees (Comité Consultatif International Téléphonique, Comité Consultatif International Télégraphique, and Comité Consultatif International Radio) as part of the Union. The new Convention strengthens the two latter (Article 8), and provides for the headquarters of all three to be at Geneva. It also creates two new permanent organs of the Union, namely, an Administrative Council and an International Frequency Registration Board.

The Administrative Council, which will consist of representatives of 18 member countries of the Union, shall normally meet at Geneva once a year and shall act in behalf of the Plenipotentiary Conference in intervals between its meetings, which are normally every five years, always within the limits of the power delegated to it by the latter. An important duty assigned to this Council is that it shall be responsible for taking all steps to facilitate the implementation by the members of the Union of the dispositions of the Convention, of the regulations, and of the decisions of the Plenipotentiary Conference.

As other duties concern the appointment of the Secretary General (and two assistants), supervision of the administrative functions of the Union, approval of its annual budget, audit of accounts, etc., it may be said that the Administrative Council, in fact, constitutes the management of the organisation of the Union.

The essential duties of the International Frequency Registration Board are to effect an orderly recording of frequency assignments made by different countries and to advise them of the maximum practicable number of radio channels that is workable without harmful interference. Members of the Board, who are all nationals of different countries, shall be elected by each ordinary administrative radio conference according to its procedure. They are to be full-time paid officials of the Union. Perhaps, the most important feature is that members of the Board shall serve, not as representatives of their respective countries or a region, but as custodians of an international public trust. Also, no member of the Board shall request or receive instructions relating to the exercise of his duties from any government, public or private organisation, or person.

The general duties of the Bureau of the Union were dealt with in Article 17 of the Madrid Convention. The corresponding provisions in the new convention (Article 9) lay down similar duties for the Secretary General. He is, however, given additional powers, such as the appointment of staff, and the relations between the Secretariat and the Administrative Council are defined.

A separate article has been provided in the Convention on the subject of Administrative Conferences, which are to be held at the same place and at the same time as the Plenipotentiary Conference, in general, every five years. The primary function of the administrative telegraph, telephone, and radio conferences will be to revise the regulations annexed to the Convention and to deal with all other matters deemed necessary within the terms of the Convention and regulations, or in accordance with any directive given by the Plenipotentiary Conference.

Provision is also made for calling extraordinary conferences during the intervals between normal conferences.

The expenses of the Union are divided into (*a*) general or ordinary expenses, which cover the day-to-day activities of the Union (including the new International Frequency Registration Board and the strengthened consultative committees) and are borne by all members and associate members, and (*b*) extraordinary expenses, which cover the costs of the various conferences and meetings of the consultative committees and are borne by the members and associate members participating. Under the old Convention, private operating agencies and international organisations participating in the work of the consultative committees also contributed to the expenses of their meetings. Under the new Convention, this system is extended to cover Administrative Conferences as well; and in addition scientific or manufacturing organisations contribute to the expenses of the meetings of the consultative committees.

In the past, the Swiss government has advanced funds for the operation of the Union and has been reimbursed when the members' contributions (paid in arrears) were received. Under the new Convention, the Union is financially autonomous, and from the start of 1949 members' contributions will be payable in advance.

French was formerly the official language of the Union, but both French and English were admitted at conferences. The new Convention makes provision for five official languages: Chinese, English, French, Russian, and Spanish. In case of dispute, the French text shall be authentic. There are also now three working languages, Spanish having been added at the instigation of the Latin American countries.

The new Convention is specific in stating that members are bound to abide by the provisions of the Convention and all three sets of regulations (i.e., the Atlantic City radio communication regulations and the telephone and telegraph regulations as amended next year at a conference to be held in Paris). Private operating agencies are likewise bound by the same rule, but military services are exempted from this obligation.

The effective date of the new Convention is January 1, 1949.

There are two annexes to this part of the Convention proper. Annex 1 contains a list of all of the countries represented at the Plenipotentiary Conference. Annex 2 presents a list of definitions of terms used in international telecommunication. Most of the definitions which appeared in an annex to the Madrid Convention have been retained and a few new ones added, due to progress made since the old Convention became effective.

1.2 General Regulations

1.2.1 Conferences

In the Madrid Convention, these provisions are to be found in an appendix, where they are grouped under the heading of rules of procedure. This, in fact, is the subject matter of the regulations. In other words, they lay down the rules to be followed in convening the plenipotentiary and administrative conferences, voting, methods of introducing proposals, order of seating, adoption of proposals, minutes and reports, etc. The regulations also include provisions for a corresponding routine, applicable to the conferences of the international consultative committees. A conference, before entering on its deliberations, adopts its own rules of procedure taking as a basis these regulations, but with such modifications as it thinks fit.

Editorially, the old regulations were listed as *Articles*, of which there were 31 in all. In the new Convention, the regulations are presented as *Rules* in chapter form. The numbers of the rules do not coincide with the former articles and, in commenting on changes or additions to the old text, reference is in all cases made to the number of the *Article.*

Generally speaking, much of the old text has been retained, but has been remodelled to include provisions for the functions of the Administrative Council.

Article 2 of the Madrid regulations specifies the method to be followed in sending out invitations to contracting governments to attend plenary sessions of the Union. In the new Convention, this article is contained in Chapter 1 of the regulations and contains two additional clauses. One states that the United Nations Organisation shall be notified of the date and place of the Plenipotentiary Conference in order that the said Organisation may attend the meeting. The second clause states that any permanent organ of the International Telecommunication Union shall be admitted, as of right, to take part in the conference in an advisory capacity.

Similar rules are laid down with respect to invitation and admission to the Administrative Conferences.

Articles 21 and 22 (Madrid) contain the rules on voting as heretofore used. In the new Convention, these articles are merged into one and a new provision introduced to the effect that each delegation shall present credentials which, in the case of plenipotentiary conferences, must be full powers signed by the head of the government or the minister of foreign affairs.

Article 11 (Madrid) states that the choice of chairmen and vice chairmen of each committee shall be ratified by the plenary assembly. The new Convention retains this clause and adds another, which empowers the chairman of each committee to propose to his committee the nomination of rapporteurs, as well as of the chairmen, vice chairmen, and rapporteurs of subcommittees. In this connection, it is to be observed that in the new rule there is no indication, except in the case of plenipotentiary conferences, that any of these officials must be administration (government) men.

2. *International Consultative Committees*

As already mentioned, the Madrid Convention proper merely states that international consultative committees may be set up. The reader is referred to the individual sets of regulations for further particulars.

The new Convention lays down general rules for participation in such committees. As these new rules are important, they are summarised below. Administrations as members of the Telecommunication Union are admitted as a right. Recognised private operating agencies are admitted at their own request, which must, however, be approved by the administration of the government recognising it. International organisations, which co-ordinate their work with the Union, may be admitted in an advisory capacity. Scientific or manufacturing organisations engaged in telecommunication are admitted in an advisory capacity to committees of the plenary assembly and to study groups, subject to the approval of the administration of the countries concerned.

Heretofore only the Comité Consultatif International Téléphonique has had a permanent secretariat in charge of a General Secretary. According to the new Convention, the telegraph and radio committees also will have permanent secretariats. All three committees will each have a director, who, in effect, will have much the same status as the Secretary-General of the telephone committee had in the past.

The general regulations conclude with a new clause on the financing of the consultative committees. The salaries of the directors of the consultative committees and the ordinary expenses of their secretariats shall be included in the ordinary expenses of the Union. The expenses of the meetings, plenary and committee, including the extraordinary expenses of the directors and those of their secretariats, shall be borne by the administrations, private operating agencies, and scientific or manufacturing organisations participating. These expenses shall be apportioned among these bodies in proportion to the number of units that the respective governments contribute to the ordinary expenses of the Union. These bodies shall also defray their own personal expenses. All documents are to be purchased.

3. *Protocols and Resolutions*

This part chiefly concerns arrangements for the transitional period, i.e., up to January 1, 1949, when the new Convention comes into force. Thus the Administrative Council was to be set up forthwith and function on a provisional basis until January 1, 1949. In fact, it held its first meeting at Atlantic City, elected its chairman and vice chairmen and arranged for its subsequent sessions. The second session was held in Geneva (January to February, 1948) and resulted in many problems in connection with the operation of the new Convention being clarified. Another meeting will be held in September, 1948, to deal with any further questions that arise before the entry into force of the new Convention. The International Frequency Registration Board has likewise been set up. It held its first meeting at Atlantic City, organised its work, and is operating on a temporary basis until January 1, 1949. The general secretariat will use the present personnel of the Berne Bureau, pending the completion of new staff arrangements.

The protocol gives particulars of the ordinary annual expenses of the Union for 1948 and for the period 1949-1952.

For 1948, the secretary general of the Union is authorised, on approval by the administrative council, to incur ordinary expenses not exceeding 1,500,000 Swiss francs. Two-thirds of this amount is to be allocated to the radio division and one-third to the telephone and telegraph divisions. In addition, the expenses of the International Frequency Registration Board and the Provisional Frequency Board during 1948 amounting to 1,832,000 Swiss francs are authorised.

The administrative council may incur ordinary annual expenses not exceeding 4,000,000 Swiss francs for the period 1949-1952. The prior approval of the majority of members and associate members of the Union is required if this sum is to be exceeded.

The resolutions concern grading of the secretarial staff, pension funds, financial, and other matters.

Regarding the payments of annual dues from member administrations and other recognised members, a resolution was passed instructing the secretary general to show in the annual financial report, beginning 1947, a list of the countries in arrears, together with the sums due.

The section containing the resolutions, concludes with an opinion that is worthy of note. It reads: "The International Telecommunication Conference of Atlantic City recognises the necessity of rendering immediate assistance to the countries (Members of the Union) that were devastated by the second world war, in order to rehabilitate their telecommunication systems, and expresses the hope that the United Nations draw attention of its competent organs to the importance and the urgency of this problem, which is part of the general problem of reconstruction."

4. *Agreement with United Nations Organisation*

This is a formal agreement between the United Nations Organisation and the International Telecommunication Union. The most important article in this agreement is the first one which reads: "The United Nations recognises the International Telecommunication Union as the specialised agency, responsible for taking such action as may be appropriate under its basic instrument for the accomplishment of the purposes set forth therein."

The other articles in the agreement cover methods and means of co-operation between the two bodies.

5. *Historical Note*

As a matter of record, it may be mentioned that the International Telecommunication Union was originally known as the International Telegraph Union, created by the International Telegraph Congress in Vienna, 1868, for the purpose of collecting, arranging, and publishing information of every kind concerning international telegraphy. An office was established in Berne, Switzerland to serve as headquarters and was known as the Bureau of the Union, which was financed by contributions from member countries and from the sale of publications. Plenary meetings of the Union were held at regular intervals of 3 to 5 years, and in Paris in 1890 telephony was for the first time brought in as part of the plenary proceedings. In 1906, the Berlin Radio Telegraphic Conference entrusted to the Bureau the same duties in regard to radiotelegraphy as it already discharged in the field of telegraphy. From January 1, 1929, the international radiotelegraph service was governed by the Washington Convention of November, 1927, and the general regulation annexed thereto.

The International Telecommunication Convention Madrid (1932) and its three annexed sets of service regulations (for international telegraphy, telephony, and radiocommunication, respectively) were the result of the fusion of the Telegraph and Radiotelegraph Conventions. The three sets of service regulations were revised at Cairo (1938).

High-Stability Radio Distance-Measuring Equipment for Aerial Navigation

By H. BUSIGNIES

Federal Telecommunication Laboratories, Incorporated, Nutley, New Jersey

DISTANCE-MEASURING equipment constitutes an important radio aid to aerial navigation. It provides an equipped aircraft with a simple meter indication of its distance from a selected ground beacon. Nonrotating antennas having omnidirectional radiation patterns in the horizontal plane are employed in the aircraft and on the ground. Line-of-sight distances up to 100 nautical miles (185 kilometers) are measured with an accuracy at slant range of ±0.2 mile or 1 percent, whichever is greater.

The crystal-controlled equipment to be described provides 51 clear two-way narrow-band channels in the 960–1215-megacycle-per-second aerial-navigation band. A special detector permits adjacent-channel rejection to exceed 70 decibels with a channel separation of only 2.375 megacycles.

A conservatively rated beacon services 50 aircraft, and several times this number may be accommodated by two or more beacons operating in parallel.

The airborne equipment, exclusive of meter, antenna, and remote control box, is housed in a standard aircraft radio case $15\frac{9}{16}$ by $10\frac{1}{8}$ by $7\frac{5}{8}$ inches (40 by 26 by 19 centimeters) and weighs less than 35 pounds (16 kilograms). The antenna is a simple half-wave element. The standard ground beacon is housed in a cabinet with a maximum height of 76 inches and will deliver 5 kilowatts of peak power. The beacon antenna is a vertical array of nine discone elements.

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The indication in an aircraft of its distance relative to a fixed ground beacon has been stated by various policy-making organizations to be the most urgently needed radio aid to aerial navigation. The combined use of omnirange beacons and distance-measuring equipments allows navigation by the distance-bearing (R-θ) system[1] recommended by the International Civil Aviation Organization. Also, it is contemplated that distance indicators may supplement or replace the marker beacons used with the instrument landing system.[2] As a result of the concentrated effort that has been applied to developing a practical distance-measuring equipment,[3] the major problems have now been solved and several equipments have been put in experimental operation.[4]

1. *Principles of Operation*

In the distance-measuring system to be described, distance is measured by an airborne challenger working in conjunction with a ground beacon. The challenger-responder principle is utilized. Both the airborne challenger and the ground beacon have a pulsed transmitter and a receiver. The transmitter in the aircraft starts the measuring process when it sends out a challenging pulse. This is received at the ground beacon and causes its transmitter to respond with a similar pulse. When the response pulse is received in the aircraft, special circuits measure the time elapsed between the transmission of the challenging pulse and the reception of the response pulse. Other circuits then convert the time difference into a mechanical indication of the distance from the aircraft to the beacon. This sequence of operations is repeated frequently enough to give a smooth and continuous indication.

A beacon does not transmit a response instantaneously when challenged; a standard delay of 75 microseconds has been adopted and all beacons are adjusted to this value. The airborne distance-indication equipment allows for this delay in its computations.

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[1] R. I. Colin, "Survey of Radio Navigational Aids," *Electrical Communication*, v. 24, pp. 219-261; June, 1947.

[2] S. Pickles, "Army Air Forces' Portable Instrument Landing System," *Electrical Communication*, v. 22, n. 3, pp. 262-294; 1945.

[3] H. Busignies, P. R. Adams, and R. I. Colin, "Aerial Navigation and Traffic Control with Navaglobe, Navar, Navaglide, and Navascreen," *Electrical Communication*, v. 23, pp. 113-143; June, 1946.

[4] P. R. Adams, S. H. M. Dodington, and J. A. Herbst, "First Tests on Navar System for Aerial Navigation and Air Traffic Control" (Summary), *Electrical Communication*, v. 24, pp. 263-264; June, 1947.

In order that many aircraft may operate in a given area with a single ground beacon, the beacon responds to all challenges in its assigned channel. Each airborne challenger, therefore, receives the responses of the beacon to other challengers. Thus, the airborne equipment must have some means of finding and abstracting the responses to its own pulses. For this purpose, special circuits called the *strobe* examine in a stroboscopic manner all responses received. Only those response pulses that have the same characteristic timing as the challenging pulses emitted from that aircraft affect its distance indicator.

The number of ground beacons to be employed within a given area will depend on the extent of the area, nature of the terrain, number of airports within it, density of traffic, and accepted flight procedures. It is generally agreed that each beacon will be limited to line-of-sight operation corresponding to a maximum range of about 100 nautical miles (185 kilometers) and approximately 40 nautical miles (74 kilometers) for an airplane at a 1000-foot (305-meter) altitude and a beacon-antenna height of 50 feet (15 meters) above mean elevation. There must be complete coverage of the airways.

2. Channeling System

Each beacon must operate on a separate and well-defined two-way channel, and each airborne challenger must provide for simple selection of any one beacon. The method of channeling used in a distance-measuring equipment to provide a distinct two-way channel for each beacon must satisfy the following basic requirements:

A. Economy in utilization of available frequency spectrum. (In the U.S.A., the number of channels was set at 51 in the frequency band from 960-1215 megacycles.)

B. Separation of channels must be sufficient to eliminate adjacent-channel interference.

C. Each beacon channel must accommodate a large number of aircraft without interference among them.

D. Desired distance range must be obtained with minimum transmitted power.

E. Distance-measuring equipment should not interfere with any other function of a complete aerial-navigation system, such as aircraft identification and altitude reporting, that might share the same channels.

These requirements, which affect the development not only of distance-measuring equipment but other future navigational aids in the 1000-megacycle region, have posed major issues in establishing specifications for a distance-measuring system.

The history of radio provides ample evidence of the continuous trend toward crystal control. Whenever a new band of frequencies is opened, there is always such demand that the most readily available circuits are immediately pressed into service without too much thought of frequency stability. Often, highly ingenious and complicated circuits and techniques are evolved to combat the basic limitations in stability. Sooner or later, however, closer frequency tolerance, with resultant better use of the spectrum and better performance of the equipment, becomes so imperative that the best methods must be adopted, regardless of practical difficulties.

It was felt, therefore, that the future of the proposed aids in the 1000-megacycle region could best be served by a direct attack aimed at the requisite number of crystal-controlled narrow-band channels. Development of such a system presented a distinct technical challenge, but one that would result in a standard type of equipment that could be built by different groups in various countries according to more or less standard production routines.

Some of the specific characteristics of a high-stability, crystal-controlled system may be seen from the following list of characteristics of equipments already constructed.

A. Channel separation of 2.375 megacycles allows 51 two-way clear channels within the allocated band of 960 to 1215 megacycles. The challenging channels cover from 965.375 to 1084.125 megacycles in 2.375-megacycle intervals. The reply channels are always 125.5 megacycles above the challenging channel and extend from 1090.875 megacycles to 1209.625 megacycles.

B. Adjacent-channel rejection is so complete (obtained through the use of a special detector) that in the geographic distribution of beacons, those that are physically adjacent may operate on adjacent channels. Channel selection is simplified and operation is free of interference. Sharing of channels with other services, such as omnirange and localizer beacons, on a geographic basis becomes a simple matter.

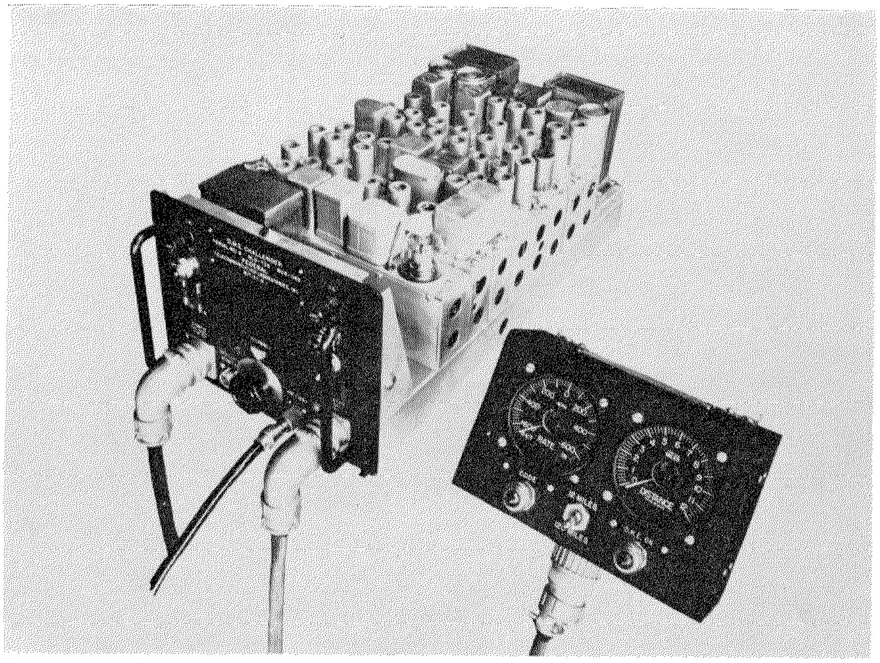

Fig. 1—Airborne distance-measuring equipment flight tested early in 1947. Indicators are provided for distance from the beacon and for the aircraft speed in miles per hour. With the exception of the indicators, all the equipment is mounted on a single chassis.

C. Narrow bandwidth greatly reduces the possibility of interference from harmonics of lower-frequency transmitters.

D. A given ground beacon can service a large number of aircraft, because high stability reduces the total number of pulses required per aircraft.

E. As pulse combinations or codes are not employed for channeling, the use of multiple pulses is reserved for other purposes, including the transmission of additional navigational or traffic-control information over the distance-measuring channels.

3. High-Stability Distance-Measuring Equipment

3.1 GENERAL

A complete distance-measuring system comprising an airborne challenger and a ground beacon has been developed. In both equipments, the receivers as well as the transmitters are crystal controlled. In each case, the receiver and transmitter, together with the necessary video-frequency, power, and measuring circuits, are housed in a single cabinet. The same antenna is used for both transmission and reception.

3.2 AIRBORNE CHALLENGER

Fig. 1 is a photograph of one of a series of airborne equipments flight tested early in the spring of 1947 to demonstrate system performance. This early series of challengers operates on only the first 6 of the 51 channels and gave distance readings on a 270-degree-scale meter. In this particular instrument, two ranges, 0–12 and 0–120 statute miles (10.4 and 104 nautical miles or 19.3 and 193 kilometers), are provided. Indication of the rate of travel of the aircraft is an optional feature and provision is made for a maximum speed of 600 miles (965 kilometers) per hour inbound and outbound.

A feature of the equipment is the subunit method of construction, with many of the subunits being of the plug-in type.

Fig. 2 is a photograph of a later equipment that operates on 51 channels. In this model, distance information is obtained directly from the rotation of a shaft and can be supplied to the computer of an automatic-flight-control equipment. The distance indicator is a panel-mounting type having a rotating disc and a rotating pointer. The calibration is from 0 to 100 nautical miles (185 kilometers), with the smallest scale division equal to 0.1 mile.

This equipment is mounted in a standard *B1–C1* aircraft radio case measuring $15\frac{9}{16}$ by $10\frac{1}{8}$ by $7\frac{5}{8}$ inches (40 by 26 by 19 centimeters) and weighs less than 35 pounds (16 kilograms). It operates from 115-volt 380–1760-cycle and 28-volt direct-current supplies, requiring 250 watts. It is operable over a range of ambient temperature from -40 to $+60$ degrees centigrade (-40 to $+140$ degrees fahrenheit) at a relative humidity varying between 5 and 95 percent and within the barometric pressure range corresponding to 8.5 and 31 inches of mercury.

A few of the circuit details may be interesting.

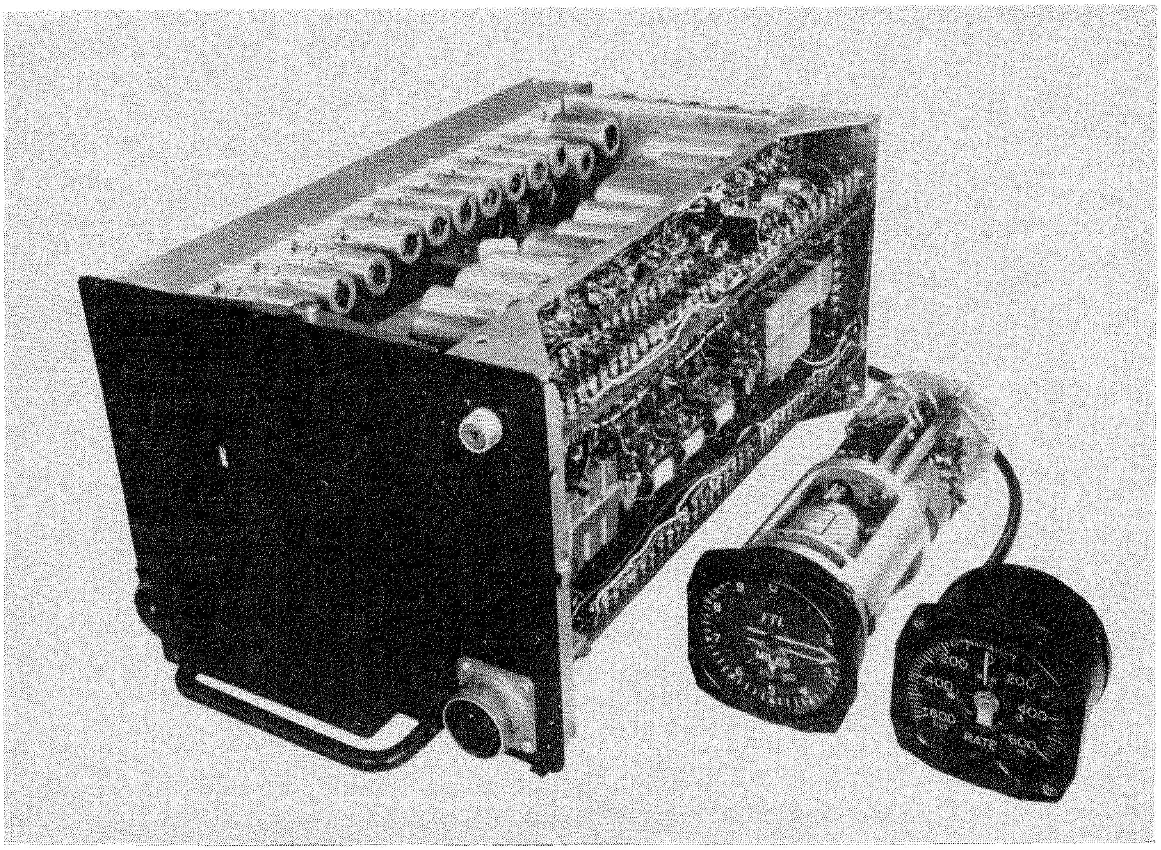

Fig. 2—Latest airborne distance-measuring equipment. Fifty-one crystal-controlled channels are provided. Distance indication is given by the rotation of a shaft. Rate indicator is optional. The remote control box is not shown.

Each of the 51 channels has two clear crystal-controlled frequencies, one for transmission and one for reception, having a constant difference of 125.5 megacycles. The two frequencies of each channel are determined by a single crystal, and 51 crystals in a rotating turret are provided for the 51 channels. The single tuning control of the crystal oscillator, the tuning plunger of the transmitting oscillator, and the tuning plunger of the receiver preselector cavities are linked to the channel-selector mechanism.

The local crystal oscillator and frequency multiplier supply a heterodyning signal to two radio-frequency converters. During transmission, a sample of the transmitting oscillator energy is compared in the automatic-frequency-control converter with the crystal-controlled local-oscillator signal; the difference frequency operates the automatic-frequency-control circuits and the tuning mechanism that brings the transmitter to the correct frequency. It is possible to hold the frequency to within ±100 kilocycles of the assigned value; but this extreme accuracy does not appear to be necessary and a tolerance of ±200 kilocycles is satisfactory. During reception, the incoming signal passes through the receiver preselector cavities, receiver radio-frequency converter, intermediate-frequency amplifier, and a special detector that is instrumental in obtaining an adjacent-channel rejection greater than 70 decibels. The video-frequency output of the detector is fed to the distance-measuring circuits and indicators.

For continuous distance indication, the receiver requires an input signal of 15 microvolts across its 50-ohm input terminals. The strobe circuit requires approximately 7 seconds to search through the distance range of zero to 100 nautical miles (185 kilometers) for the response from the beacon. It will continue to search until a response is received and will then lock onto the signal from the beacon and give a steady indication of distance. The accuracy of indication is within ±0.2 mile or 1 percent, whichever is greater, under normal slant-range conditions.

When a continuous indication of distance is being obtained, the strobe circuits are locked to the responses from the beacon and the airborne challenger emits approximately 30 pulses per second. When no distance indication is being obtained, the strobe circuits search through the distance range for responses from the desired beacon, and the pulse-repetition rate is increased to 150 pulses per second. For a pulse length of 1.5 microseconds, the duty cycles for tracking and searching are $30 \times 1.5 \times 10^{-6}$ and $150 \times 1.5 \times 10^{-6}$, or 0.0045 percent and 0.023 percent, respectively. With a maximum allowable duty cycle of 0.1 percent, the duty cycle during search is only one-fifth of the maximum allowable; and during normal tracking operation there is a much greater reserve of power. In the future, it is planned to utilize this available power to transmit to the ground such information as the identity and altitude of the aircraft.

The antenna serves for both transmission and reception and is mounted on the belly of the airplane. The standard antenna installation employing 35 feet of RG-8/U radio-frequency coaxial cable, has a loss in the cable under matched conditions of 3 decibels in each direction. The peak power of the transmitting oscillator is 2 kilowatts nominal; the peak power actually radiated by the antenna, therefore, is nominally 1 kilowatt.

3.3 Ground Beacon

The function of the ground beacon is to respond to challenges from airborne equipments operating on the frequency assigned to the beacon. However, in a comprehensive system for aerial navigation and traffic control, other functions may be superimposed on the distance-measuring beacon to transmit information to the aircraft.

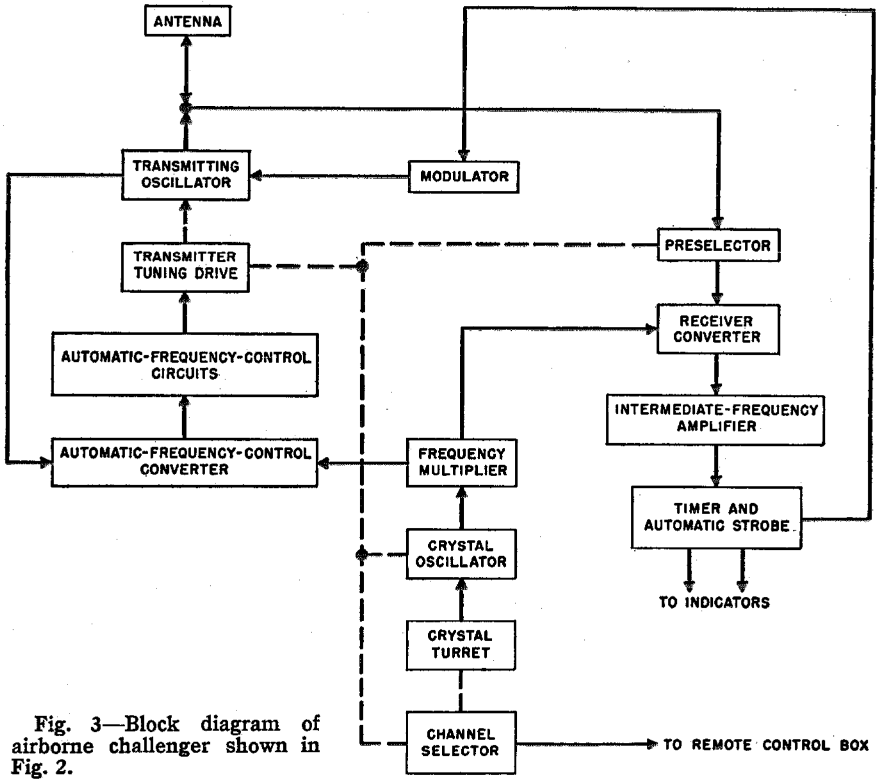

Fig. 3—Block diagram of airborne challenger shown in Fig. 2.

The ground beacon, shown in Fig. 4, may be tuned to any one of the 51 channels assigned to this service. It has a nominal peak-power output of 5 kilowatts and minimum and maximum outputs of 3.4 and 6 kilowatts, respectively. It requires less than 1000 watts from a 115-volt 50–70-cycle single-phase power source. Using standard 19-inch (48-centimeter) panels, it is 76 inches (193 centimeters) high and weighs less than 500 pounds (227 kilograms). It is designed for operation between −40 and +60 degrees centigrade (−40 and +140 degrees fahrenheit), 5- and 95-percent relative humidity, and atmospheric pressures equivalent to 17 to 31 inches of mercury.

The beacon features simplified construction with subchassis mounted vertically to provide ready access to the wiring and test points from the front of the cabinet and to the tubes and large components from the rear.

Operationally, these beacons may be used for general navigation or for approaching an airport. For navigation, combined use is made of an omnirange beacon and distance-measuring equipment. A maximum range of 100 nautical miles (185 kilometers) has been established as being satisfactory for this purpose. For the next few years, the density of air traffic probably will be such that not more than 50 aircraft will be within the service range of a distance-measuring beacon, i.e., within a radius of 100 nautical miles. Accordingly, the present beacon has been designed to handle 50 aircraft with a superior service and up to 80 aircraft with acceptable service.

The traffic-handling capacity of a beacon is determined by the ability of the transmitting tubes to dissipate power appearing in the form of heat. Depending on the tube and circuit employed, the maximum allowable duty cycle may fall somewhere between 0.1 and 1 percent. With a total of 50 aircraft, one of the worst conditions is represented when 5 are searching and 45 are tracking. At this instant, 100-percent replies by the beacon would require a nominal duty cycle of 0.315 percent. However, operation with 80 percent replies by the beacon is considered to be adequate, and the present beacons have a maximum duty cycle rating of 0.25 percent. In the near future, the duty-cycle rating will be increased to 0.4 percent to cover functions other than distance measuring.

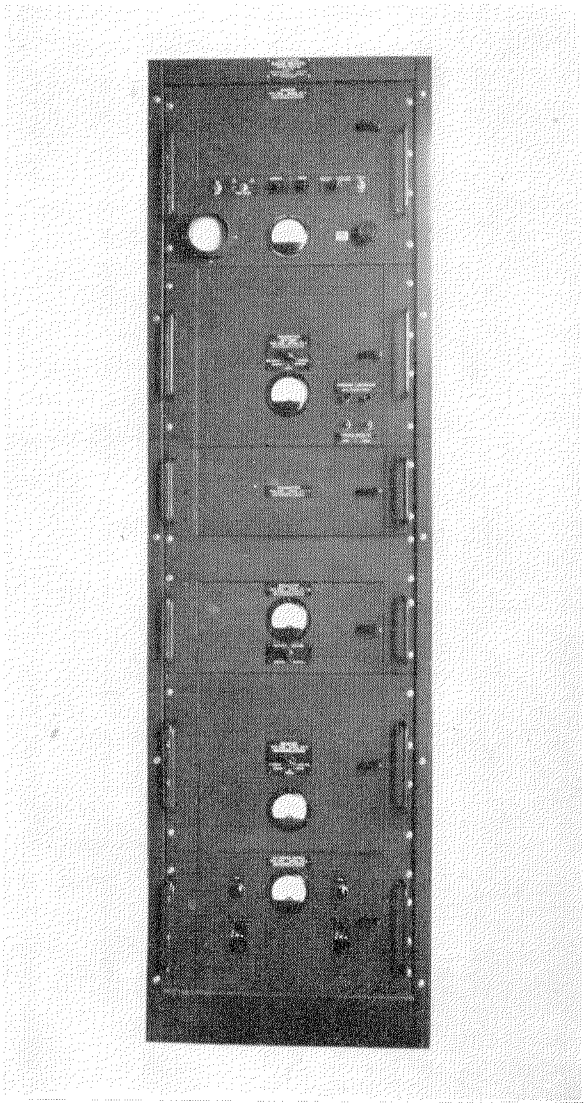

Fig. 4—Ground beacon for distance measuring. The transmitter is rated at 5 kilowatts nominal and 6 kilowatts maximum peak power. It may be adjusted to any of the 51 channels assigned to this service.

The transmitter will be triggered by any input to the 50-ohm receiver terminals exceeding 15 microvolts. The adjacent-channel rejection of the receiver is greater than 75 decibels and the image and intermediate-frequency rejections are greater than 70 and 80 decibels, respectively.

Flight tests have demonstrated that the high-stability distance-measuring equipment performs satisfactorily when less than 2 kilowatts of peak power are radiated from the ground. Therefore, a beacon rated at a peak power of 5 kilowatts is adequate for all practical installations and permits the antenna to be as far as 70 feet from the transmitter-receiver. In this maximum case, slightly more than half the total power is lost in the radio-frequency transmission line.

It should be noted that the beacon requires an antenna having an omnidirectional radiation pattern in the horizontal plane, and a low-angle beam approximately 10 degrees wide in the vertical plane. Thus, aircraft at large angles of elevation receive enough stray radiation for correct operation, but maximum power is radiated near the horizon where it is most needed. The antenna should be at least 50 feet (15 meters) above the mean elevation and should be connected to the equipment by a maximum of 70 feet (21 meters) of RG-14/U cable.

Fig. 5 is a photograph of one of the series of antennas[5] that have been built and operated according to this specification. In this installation, a single 2-kilowatt beacon is housed in a weatherproof case and mounted directly on the mast. This reduces the transmission line to about 15 feet (4.6 meters) and minimizes losses in the cable. The experimental 2-kilowatt beacon has been used for many flight tests and demonstrations, and the antenna has been used with a remote (60 feet, 18 meters) standard beacon for many other tests.

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[5] A. G. Kandoian, W. Sichak, and R. A. Felsenheld, "High Gain With Discone Antennas," *Proceedings of the National Electronics Conference, Chicago, Illinois, November 3-5, 1947*, pp. 336-346; 1947: and *Electrical Communication*, v. 25, pp. 139-147; June, 1948.

4. *Flight Tests*

From June, 1946, to October, 1947, the distance-measuring system was subjected to 130 hours of flight testing and was demonstrated on 94 different flights. The majority of the flights were made in the Laboratories' airplane, a Douglas DC-3, but a few were made in another aircraft of the same type. On many of the flights, azimuth-indicating and distance-measuring equipments were used to illustrate aerial navigation by the R-θ system.

Ground beacon transmitters of three different power output capacities have been used. These transmitters have been rated at peak-power outputs of 8, 5, and 2 kilowatts. The transmission-line losses have reduced the radiated powers to approximately 4, 2.5, and 1.7 kilowatts, respectively. All three transmitters have given equally good service out to maximum range. Therefore, for radiated powers greater than 1.7 kilowatts, the maximum range has been determined by the line-of-sight propagation conditions rather than the power rating of the beacon.

The greater part of the flight time was spent at distances less than 70 nautical miles (130 kilometers), for economy and convenience in operation of the aircraft. However, many flights were made around all beacons to distances between 90 and 120 nautical miles (167 and 223 kilometers), at altitudes near 10,000 feet (3048 meters).

The antenna, the 20-foot (6-meter) mast, and the 2-kilowatt beacon are air transportable and are easily set up for use. The installation shown in Fig. 5 was made at several locations, at distances from 8 to 75 miles (15 to 139 kilometers) from a more powerful beacon operating on the adjacent channel, a separation of 2.375 megacycles. Operation with either beacon was free of interference from the other, even when the aircraft was flying in the immediate vicinity of one of them.

5. *Acknowledgment*

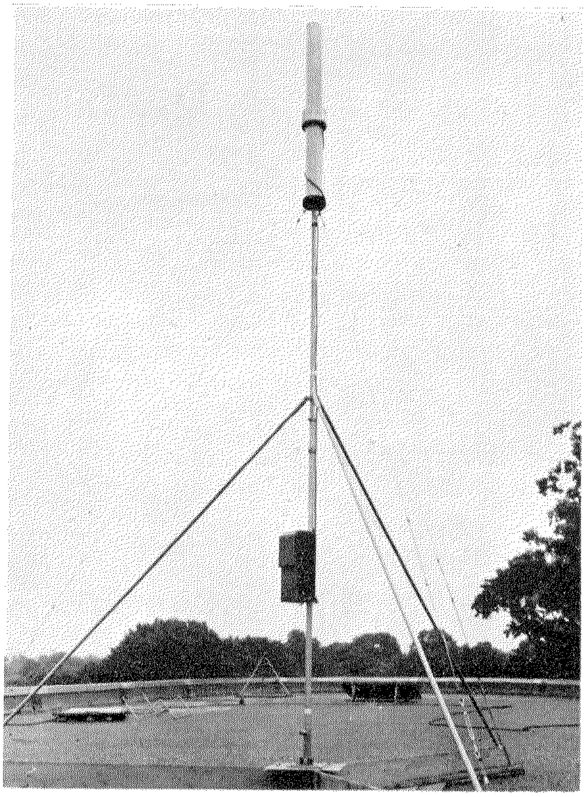

Fig. 5—Omnidirectional horizontal radiation and a narrow vertical pattern about 10 degrees wide are obtained with this stacked discone antenna. Maximum radiation is to the horizon. The air-transportable experimental beacon equipment is mounted on the mast to reduce transmission-line losses.

This development has occupied a large part of the Laboratories' efforts during the past three years, more than fifty engineers and technicians having at one time or another been involved. The basic system requirements were laid down by Paul R. Adams and the author, and have been modified from time to time by customer requirements. Over-all supervision of equipment design, ground and air, rested with Sven H. Dodington, who also designed the channeling system and the automatic frequency control. The distance-measuring circuits were designed by Gilbert R. Clark and Etienne deFaymoreau, who were greatly assisted by consultations with Norman F. Moody. The beacon antenna was designed by a group headed by A. G. Kandoian. The design of most of the mechanical details was supervised by John A. Herbst. Substantial contributions were made by Edgar W. Blaisdell, mechanical design; Warren A. Anderson, airborne intermediate-frequency amplifier and crystal multiplier; Ben Warriner, transmitters; Alex Hegedus, beacon circuitry; and Sidney Frankel, S- and X-band rejection filters. The manuscript was prepared by Henry D. Scarborough.

Aircraft Radio Communication Set A.R.I. 5272

By E. C. FIELDING

Standard Telephones and Cables, Limited, London, England

VERY-HIGH-FREQUENCY aircraft transmitter-receiver equipment, A.R.I. 5272, was developed during the war for use in naval aircraft, and is noteworthy for the economies effected in weight and dimensions relative to earlier equipments of comparable performance. Information is given regarding the various techniques and special designs used to make these economies possible.

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1. *Introduction*

The A.R.I. 5272 aircraft radio communication equipment, shown in Fig. 1, was designed primarily for use by naval carrier-based aircraft, and is constructed in accordance with Air Ministry specifications. It is, therefore, also suitable for civil flying under the most severe climatic conditions likely to be encountered, including use at altitudes up to 40,000 feet. The rotary transformer is designed for continuous operation in a horizontal position but may be used in any other attitude for short periods of time. The set is suitable for use under tropical and arctic conditions. A commercial model, designated the S.T.R.-9, has been developed.

The A.R.I. 5272 set comprises a very-high-frequency radio-telephone transmitter-receiver built in the form of a single main unit including power-supply equipment and voltage stabiliser (for valve-heater supplies), and is arranged for rapid selection of any one of four channels by means of a small remote-control unit located at a convenient point in the aircraft. The main unit is 7.9 inches high by 9.0 inches wide by 16.25 inches long (20 by 23 by 41 centimetres); the dimensions of the remote-control unit are 3.5 by 2.2 by 2.6 inches (9 by 6 by 7 centimetres); total weight is 22 pounds (10 kilograms).

Receiver and transmitter operate on the same channel frequencies and use a common aerial. The four crystal-controlled frequencies are in the range of 115 to 145 megacycles per second with frequency stability within ±0.01 per cent. The transmitter power output is 3.5 watts. Total

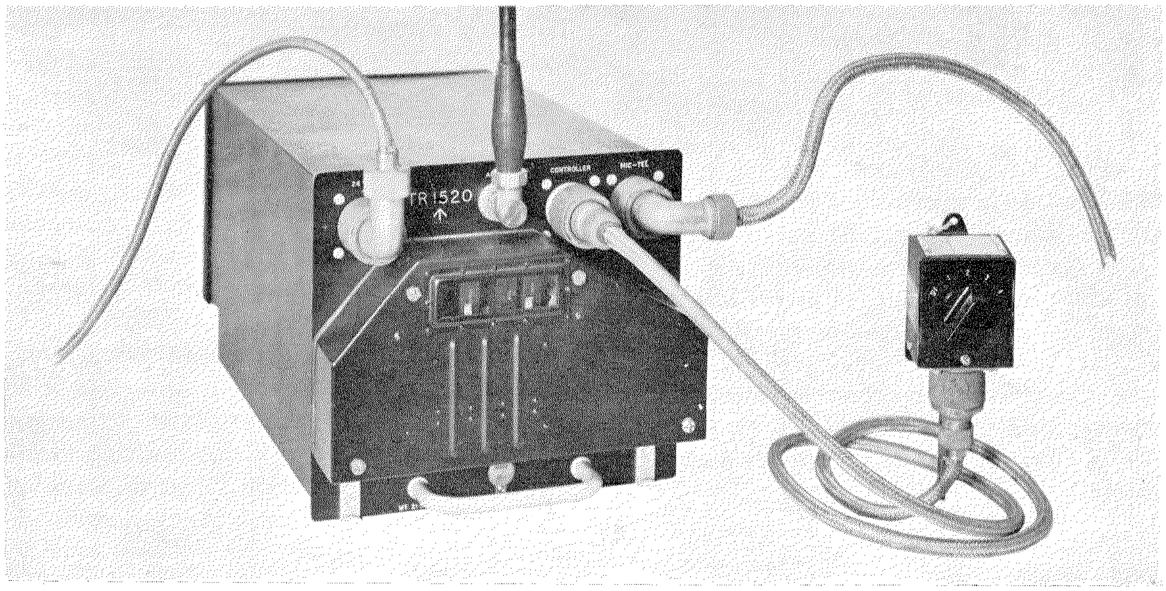

Fig. 1—A.R.I. 5272 aircraft radio communication set. The control unit is at the right. The battery cable is at the left of the main unit, the antenna is central, and the microphone-telephone cable is at the right.

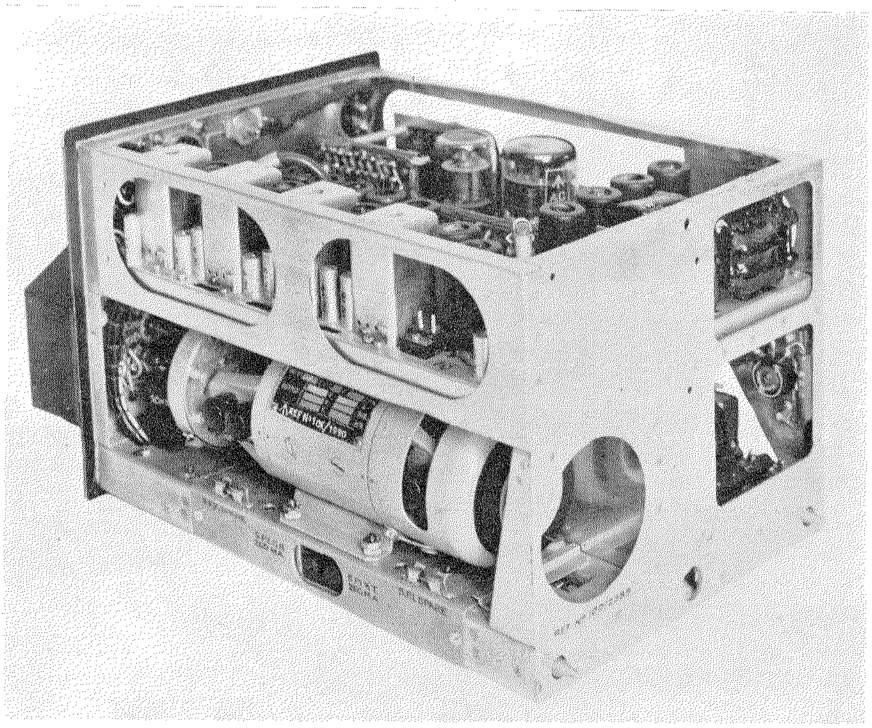

Fig. 2—Main unit showing arrangement of subsidiary assemblies. The transmitter is at the right and the receiver is at the left. The rotary transformer is in the lower left.

power requirements are 150 and 180 watts at 26 volts, the lower value being for receiver operation. The service range of the set for aircraft at a 10,000-foot altitude is approximately 200 miles between aircraft, and 100 miles from aircraft to ground stations.

The entire frequency range is available to each of the four channels, which can be set or reset independently of each other. One crystal per channel is used for both transmitter and receiver, but separate tuning adjustments are provided. During channel-changing operations, the transmitter and receiver tuning circuits are positioned simultaneously, the maximum time for this operation being a little over 3 seconds.

When the receiver is operating, the transmitter is in the standby condition with valve heaters energised; the transmitter is turned on by electrical switching of high-voltage supplies and antenna and other connections, all under control of the remote operator. Aerial switching is internal to the set and is included in the change-over mentioned above.

Gang-tuned circuits are used in the receiver and transmitter for simplicity and speed of operation. This arrangement has the added advantage that there is no risk of valves being damaged during tuning operations as a result of temporary misalignment of the different stages.

The electrical circuits are generally of a straightforward nature except that certain frequency-multiplying stages are common to both receiver and transmitter, as described later.

Valve heaters are wired in a series-parallel network to minimise the effect of the failure of any particular heater on the remainder.

2. *General Description*

The main-unit chassis, which may be seen in Fig. 2, is fitted with a detachable dust cover and is arranged for mounting in a shock-absorbing tray. It is designed fundamentally as a mounting for the subsidiary units comprising transmitter, receiver, intermediate-frequency amplifier, modulator, and power-supply unit, together with such miscellaneous components as relays, carbon-pile voltage regulator, cable form, and channel-selection mechanism. All of these subsidiary units are detachable, and are connected to the main unit by plugs and sockets. The channel-selection mechanism and all external connections are accessible at the front panel, as can be seen from Fig. 3. As might be expected in a naval airborne equipment, the fundamental consideration governing the general design, apart from the electrical requirements, was the necessity of keeping weight and size to an absolute minimum, and it will be of interest to consider in detail some of the means by which the desired results were obtained.

2.1 Unit Construction

One important reduction of size and weight has been obtained by the adoption of unit construction. If this had not been done, it would have been impracticable either to fit initially or to obtain subsequent access to large numbers of components that are completely masked by adjacent units when the chassis is fully equipped, and it would not, therefore, have been possible to make efficient use of the space within the main unit; the set would then have been larger and heavier.

Also, it was known that unit construction would greatly facilitate the manufacturing processes and enable associated groups of circuits to be bench-tested individually before fitting into the complete equipment. Similar considerations in regard to bench-testing also apply to a large extent to maintenance procedures.

2.2 Choice of Materials

Wherever possible throughout the equipment, aluminium alloys have been used in the construction of chassis and mechanisms. Manganese-aluminum alloy is employed mainly for chassis work; this being a material that, in addition to being practical from the forming and welding viewpoints, also possesses good resistance to corrosion by sea air. In cases where rigidity and wear are the deciding factors, duralumin has been used instead of the alternative heavier metals, but the latter could not be avoided in some instances for electrical or other reasons.

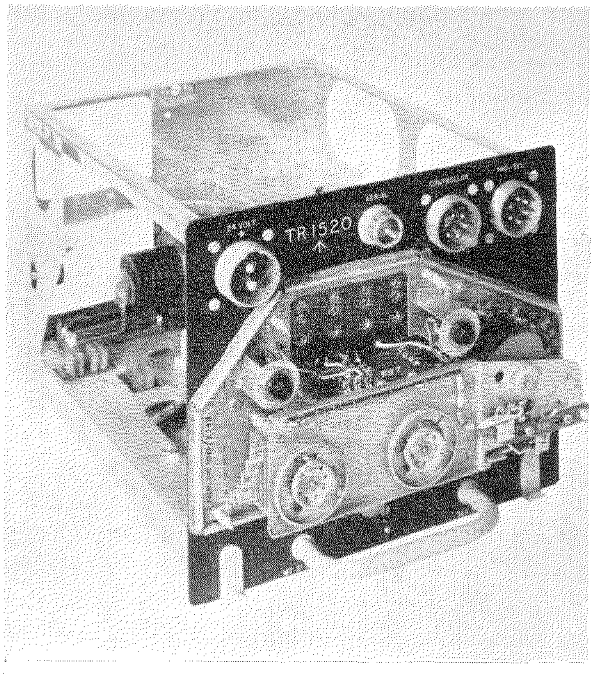

Fig. 3—Main-unit chassis showing channel-selection mechanism.

At the time that the equipment was being developed, the technique of silver-plating aluminium was insufficiently advanced, otherwise further economies could have been made in this direction.

Particular care has been exercised where materials and finishes are concerned, to avoid contacts between dissimilar metals that might lead to electrolytic corrosion.

2.3 Miniaturisation of Components

Practically all components and their mountings—including valves, plugs and sockets, and carbon-pile regulator—are of miniature types, and this represents a considerable contribution to the overall saving in weight and dimensions.

Although it is not possible to give specific figures, it is estimated that the use of miniature components has resulted in a saving of something like 50 per cent in weight and possibly more in volume relative to non-miniature components. In the case of the carbon-pile voltage regulator alone, the saving is substantial, relative figures being 10 ounces against 32, and 6 cubic inches against 26.

It might be mentioned that miniaturisation also introduced certain problems such as congestion of wiring and soldering points, which called for a considerable amount of ingenuity in their solution.

2.4 Midget Relay

The midget relay is of unusual interest, having been designed to meet the requirement for the smallest relay that would operate a maximum of two contacts with complete reliability and a standard of mechanical adjustment similar to that generally adopted on telephone switching systems. The weight of 1.5 ounces (43 grammes) and overall dimensions of 1.38 by 1 by 0.63 inches (3.4 by 2.5 by 1.6 centimetres) are considered to be the minimum possible when the variations in energising voltage and ambient

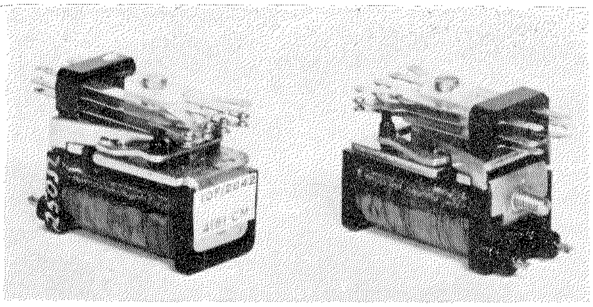

Fig. 4—Miniature relay that weighs only 1.5 ounces.

temperatures experienced in aircraft are considered. In its ordinary form, it is fitted with two sets of armature springs, each fitted with twin platinum contacts capable of handling 2 amperes continuously, or starting currents up to 10 amperes. For the latter use, a heavy-current holding winding is added to the relay to guard against contact burning, which might arise from switching the relay off whilst the rotary transformer is drawing a heavy current during starting. This winding holds the contacts closed until the current value falls to a safe figure.

For heavier duties, up to 100 amperes starting current, this type of relay can be fitted with single heavy-duty molybdenum-silver contacts. The normal winding dissipation is approximately 1.75 watts. Not least of the outstanding features of this relay is its satisfactory operation under severe vibration conditions, this being made possible by its balanced-armature construction.

2.5 Stages Common to Both Receiver and Transmitter

In view of the fact that the receiver and transmitter operate on the same frequencies, and that both are never in actual operation simultaneously, it was decided to take advantage of these conditions by making certain oscillator and frequency-multiplying stages of the receiver serve a similar function for the transmitter, thereby simplifying the circuit arrangements and saving equipment.

This feature is evident from the schematic diagram of Fig. 5 and enables both receiver and transmitter to make use of a single set of four crystals, at the expense of a single auxiliary crystal oscillator in the transmitter.

2.6 Ganged Tuning Capacitors

Similar tuning capacitors are used with slight variations for both transmitter and receiver.

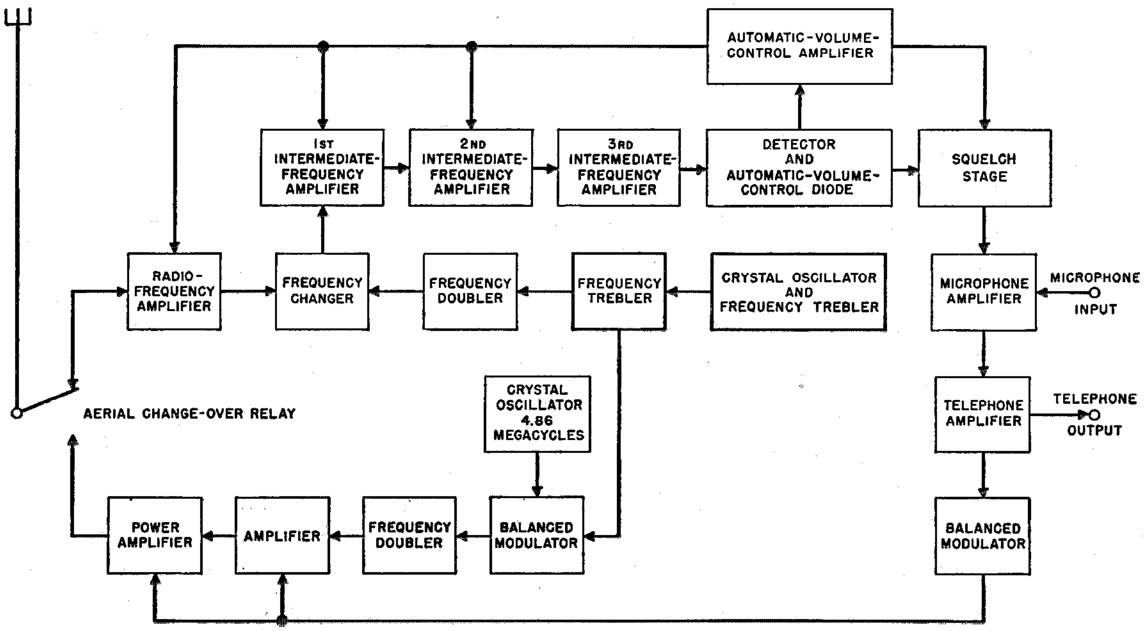
AUTOMATIC-VOLUME-CONTROL AMPLIFIER
1ST INTERMEDIATE-FREQUENCY AMPLIFIER
2ND INTERMEDIATE-FREQUENCY AMPLIFIER
3RD INTERMEDIATE-FREQUENCY AMPLIFIER
DETECTOR AND AUTOMATIC-VOLUME-CONTROL DIODE
SQUELCH STAGE
RADIO-FREQUENCY AMPLIFIER
FREQUENCY CHANGER
FREQUENCY DOUBLER
FREQUENCY TREBLER
CRYSTAL OSCILLATOR AND FREQUENCY TREBLER
MICROPHONE AMPLIFIER
MICROPHONE INPUT
AERIAL CHANGE-OVER RELAY
CRYSTAL OSCILLATOR 4.86 MEGACYCLES
TELEPHONE AMPLIFIER
TELEPHONE OUTPUT
POWER AMPLIFIER
AMPLIFIER
FREQUENCY DOUBLER
BALANCED MODULATOR
BALANCED MODULATOR

Fig. 5—Schematic diagram of equipment. The crystal oscillator and two trebler stages are common to both transmitter and receiver as are the microphone and telephone amplifiers.

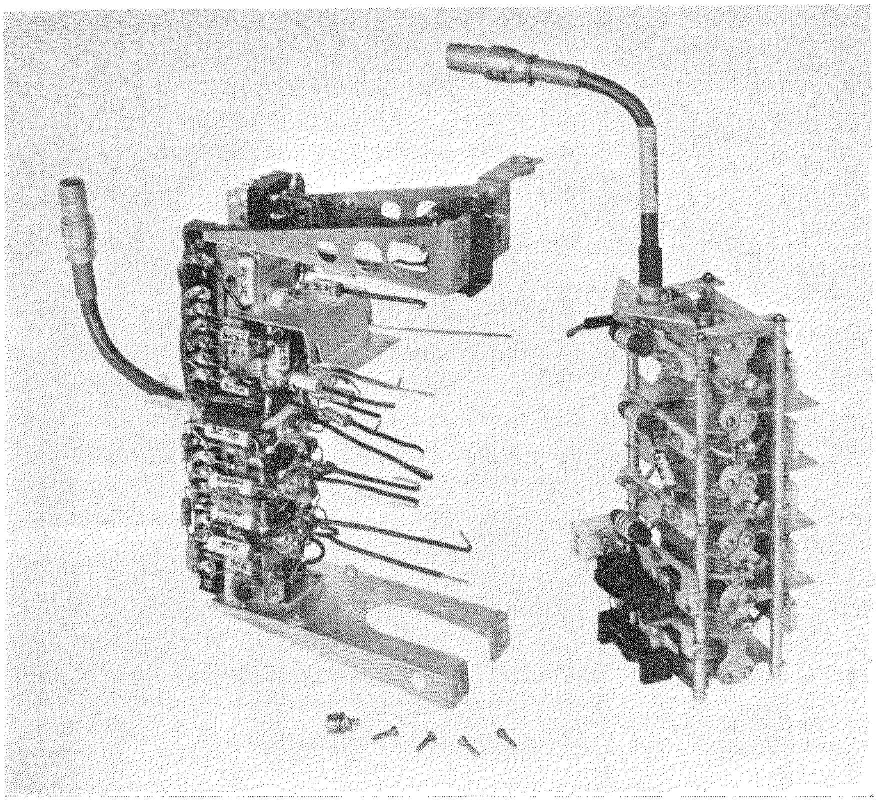

Fig. 6—Receiver input unit with 5-gang capacitors detached. The left-hand cable goes to the intermediate-frequency amplifier and the antenna input is to the coaxial cable connected to the capacitors.

They were specially designed for the equipment and, by means of compact design and silvered-ceramic technique, it has been possible to produce a unit of high electrical efficiency and small dimensions and weight, capable of being jig-assembled on a production basis. The capacitors are fitted with spring-loaded ball bearings, ensuring light action without end play, and are flexibly mounted in their respective units to avoid physical distortion. Provision is made in the design for mounting the tuning coils directly on the capacitors, thus ensuring short leads and minimum stray capacitances. Trimmer capacitors and interstage screens are built in, the latter being subsequently soldered to form integral parts of the associated screens on the unit chassis.

The complete 5-gang capacitor assembly for the receiver is visible in Fig. 6. It is 5.5 inches long by 2 inches wide by 1.75 inches deep overall (14 by 5 by 4.4 centimetres), and weighs 11 ounces (312 grammes) complete with trimmers, coils, and screens.

2.7 ROTARY TRANSFORMER

It was realised early in the development, that the considerable heat dissipation (120 watts) of the various portions of the equipment within such a comparatively small unit would inevitably require some form of forced-draught cooling system to prevent excessive temperature rise, and it was logical to make use of the rotary transformer of the power-supply unit to accommodate the blower, instead of using a separate motor for this purpose.

The presence of the blower on the rotary transformer naturally suggested the possibility of reducing the size and weight of the machine itself by the expedient of directing a proportion of the air flow through the armature tunnel to cool the winding and commutator, thus enabling a small machine to be run at an output far in excess of its normal rating. Special attention to the design of the machine end-frames and to the provision of maximum airspace in the tunnel made this a practicable proposition, with the result that instead of using an un-cooled machine 3.25 inches in diameter by 8 inches long (8 by 20 centimetres) weighing 7.5 pounds (3.4 kilogrammes), it was possible to use a machine only 2.875 inches in diameter and 8 inches long (7 by 20 centimetres) (including blower), weighing only 5.75 pounds (2.6 kilogrammes). This machine is shown in Fig. 7.

2.8 DRIVE UNIT FOR CHANNEL-CHANGING MECHANISM

In earlier equipments, it has been usual to provide a separate electromagnetic or motor-driven source of power for operating channel-switching or tuning mechanisms and these have almost invariably been both heavy and bulky.

In developing the mechanical drive unit for the A.R.I. 5272 set, a considerable proportion of the usual bulk and weight was eliminated by making still further use of the rotary transformer as the source of power. A small cam is fitted to an extension of the machine spindle, and this operates a ratchet-type drive unit fixed to the machine end-frame. The unit, whose dimensions are 2.875 inches in diameter by 0.75 inch deep (7.3 by 1.9 centimetres) over its cover, has a rotational stepdown ratio of 480:1 from the machine spindle, and a torque output of 1800 gramme-inches at approximately 15 revolutions per minute, with a total weight of only 4.5 ounces (128 grammes). The end of the machine carrying this mechanism may be seen in Fig. 8.

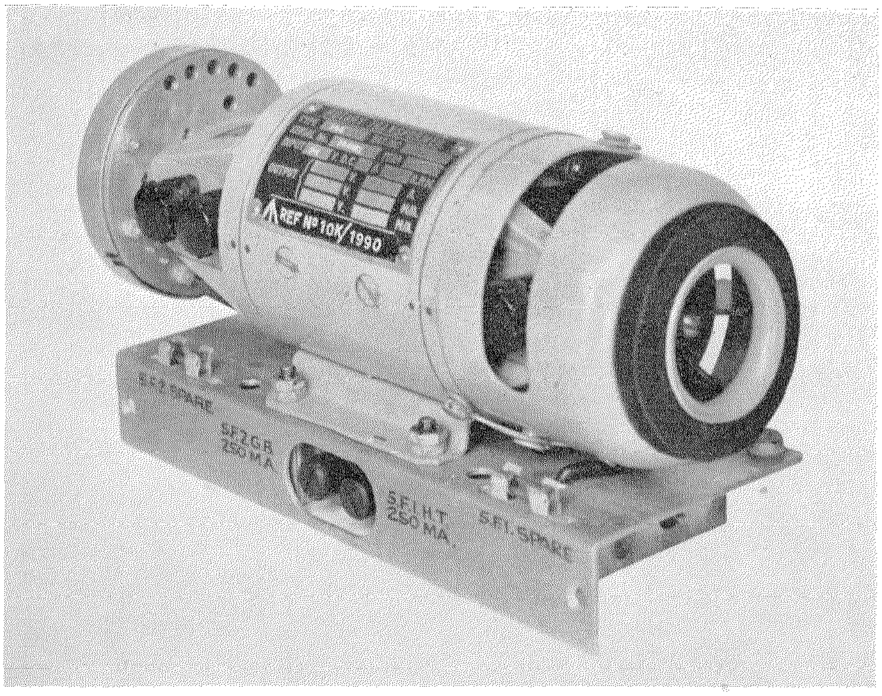

Fig. 7—Power-supply unit consisting of a rotary transformer mounted on a filter box. The drive mechanism for the channel-selecting switches is at the far end. The air blower is at the near end.

Starting and stopping is virtually instantaneous, being controlled by a small electro-magnet built into the unit. The design is such that no load, either electrical or mechanical, is taken by this drive unit except when channel changing is in progress, and the load under these conditions approximates only a 6- or 7-watt input to the set. The majority of this load is taken directly by the magnet coil, and only the relatively small remainder represents load on the rotary transformer, which is in any case operating under light-load conditions during channel-changing.

Stopping accuracy is of the order of 1.5 degrees of rotation of the output shaft, although this is actually of no considerable importance, since the channel-selection mechanism is designed to avoid the necessity for accurate

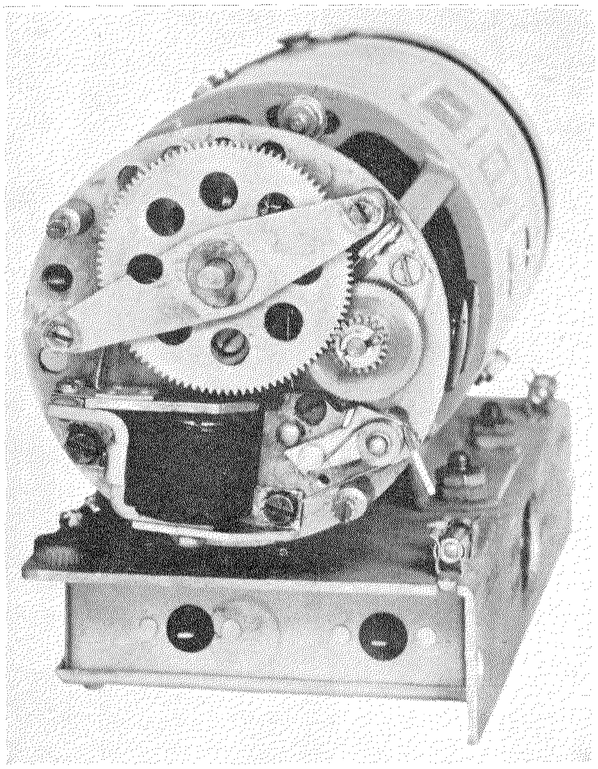

Fig. 8—Drive unit for channel selector with cover removed.

positioning of the driving shaft. When the drive unit is stopped, its output shaft is automatically locked to prevent further movement, but a release lever is provided so that it may be rotated manually for assembly and maintenance.

3. *Detailed Description*

3.1 MAIN CHASSIS

The main chassis is constructed of formed and welded aluminium alloy and is fitted with a totally enclosing dust cover of the same material.

Where silver-plated subsidiary units mount directly on the aluminium chassis, tinned surfaces are provided over the contact area to avoid undue electrolytic potentials between the metals.

The chassis layout accommodates the several subsidiary units in convenient relation to one another so that all direct interconnections are kept to minimum length. For example, the link between the common frequency-multiplying stages in the receiver and transmitter consists of knife contacts on the receiver unit that engage directly with clips on the adjacent transmitter unit.

A separate compartment is provided for the power-supply unit, with suitably positioned holes in the body of the equipment to direct the airflow to the hot spots of the set. The air blower fitted to the rotary transformer is capable of circulating 20 cubic feet (0.57 cubic metre) of air per minute through the set with a temperature rise between inlet and outlet of only 15 degrees centigrade. Further cooling of the set is assisted by the use of a matt black finish on valve cans and on the dust covers.

To prevent the access of dust and particles of foreign matter to parts of the equipment where it might have deleterious effects, air is taken in through a Vokes dry-type filter, shown in Fig. 9, and is expelled through a similar filter at the outlet. These filters are of high efficiency and low pressure drop, the design offering a large filtering surface permitting low air velocity. Particles of dust are not forced into the filter elements but rest on the surface, whence the majority are dislodged by vibration and shaken to a well at the bottom. Cleaning involves no more than removal of the filters and tapping out of the accumulated dust.

The two filters are fitted in an external compartment at the rear of the chassis dust cover, and are accessible for replacement or cleaning by removing the detachable lid. The latter is louvred for inlet and outlet of the air, and direct circulation of air between the two sets of louvres is minimised by facing them away from one another. An incidental but nevertheless valuable feature associated with the forced-air cooling system is that it promotes rapid drying out of the interior of the set after prolonged exposure to humidity.

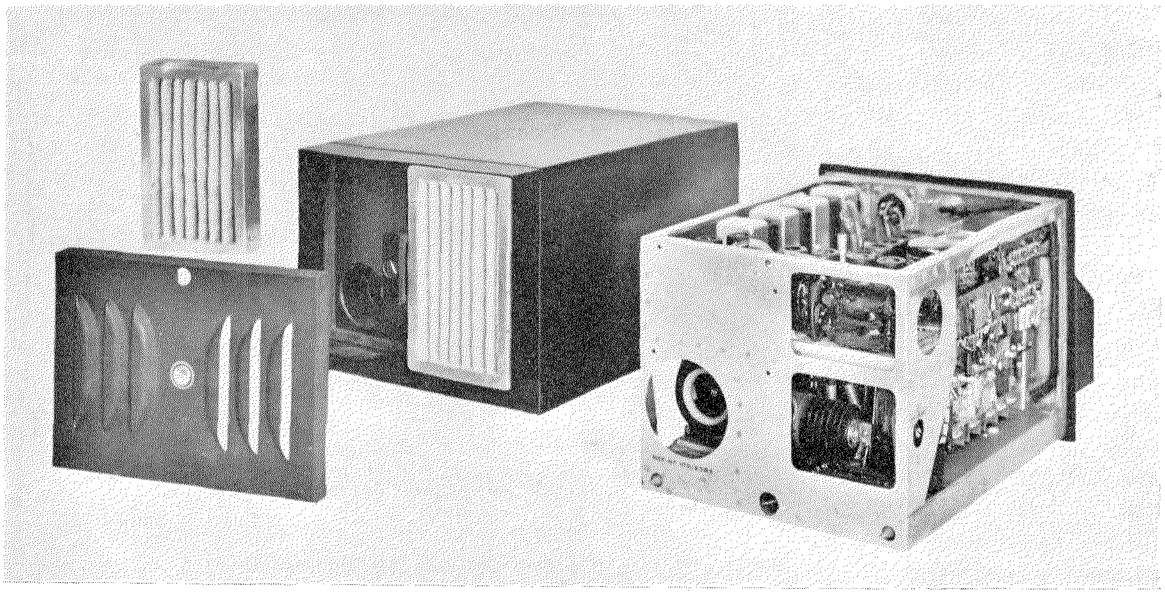

Fig. 9—Rear view of main unit showing dust cover and air intake filters.

3.2 Receiver

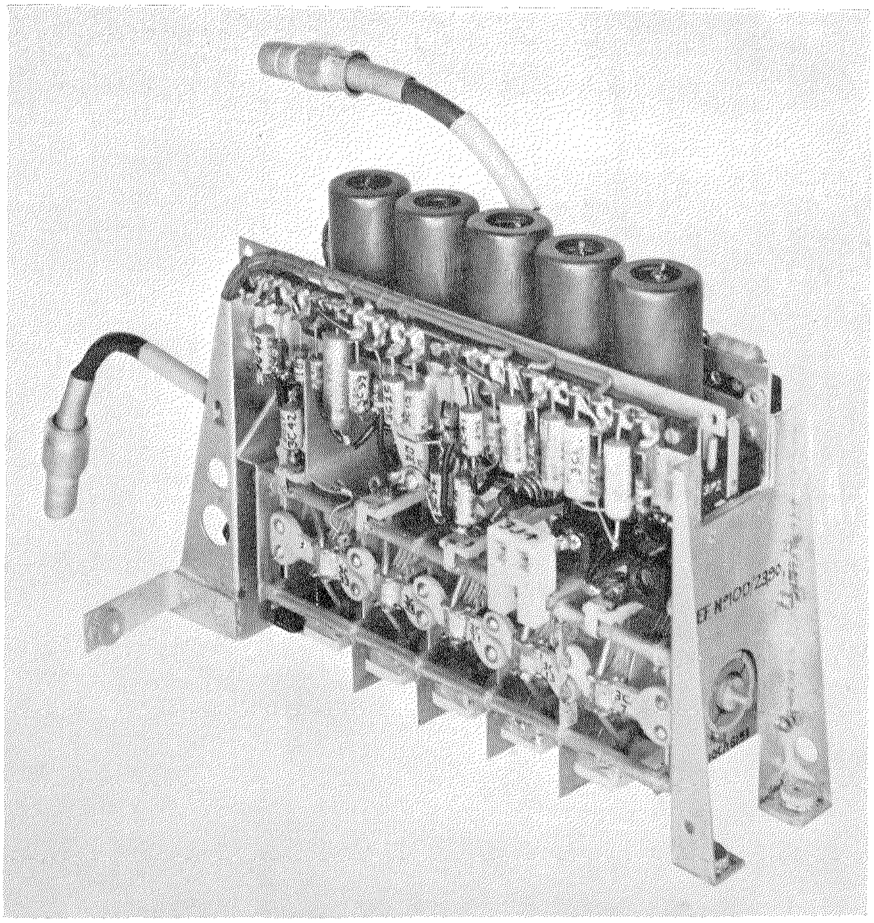

Fig. 10—Receiver input circuits with (upper) coaxial cable to intermediate-frequency amplifier.

The receiver will produce an output of not less than 20 milliwatts into a 50-ohm impedance with an input of 10 microvolts, modulated 30 per cent at 1000 cycles. The signal-to-noise ratio will be at least 10 decibels. The intermediate-frequency amplifier operates at 9.72 megacycles and has a bandwidth of not less than ±40 kilocycles at 6 decibels down from maximum response and not more than ±140 kilocycles at 40 decibels down. Suppression of second-channel signals by the receiver is not less than 35 decibels.

An audio-frequency output of 150 milliwatts will be delivered to three pairs of headphones and the equipment will operate effectively into an impedance between 50 and 150 ohms. The automatic gain control will limit the rise in output to not more than 3 decibels for an increase in input of 80 decibels above 10 microvolts. Relative to response at 1000 cycles, the audio-frequency output is between zero and −6 decibels for 300 cycles and between zero and −8 decibels for 3000 cycles.

The electrical circuit of the receiver comprises an oscillator and a trebler stage (with fundamental input frequencies controlled by four crystals), followed by another trebler and a doubler, the output of which is passed to a frequency changer where it is mixed with the incoming signal from a radio-frequency-amplifier stage. The resulting signal at 9.72 megacycles is passed via a coaxial line to the intermediate-frequency amplifier unit, which includes the detector, and thence via an audio-frequency amplifier in the modulator unit to the telephone output.

The oscillator and the two trebler stages of the receiver are also used to feed the transmitter with an input of nine times the crystal frequency.

The receiver is built in unit form on a silver-plated brass chassis, together with all valves and other components, including the 5-gang capacitor previously described. This capacitor tunes all stages simultaneously and has provision for connection to the channel-selection mechanism by means of a flexible coupling. Coil trimming is effected by dust-iron cores. These input tuning circuits of the receiver are shown in Fig. 10.

3.3 Transmitter

The transmitter will deliver 3.5 watts into a 45-ohm impedance. The carrier may be amplitude

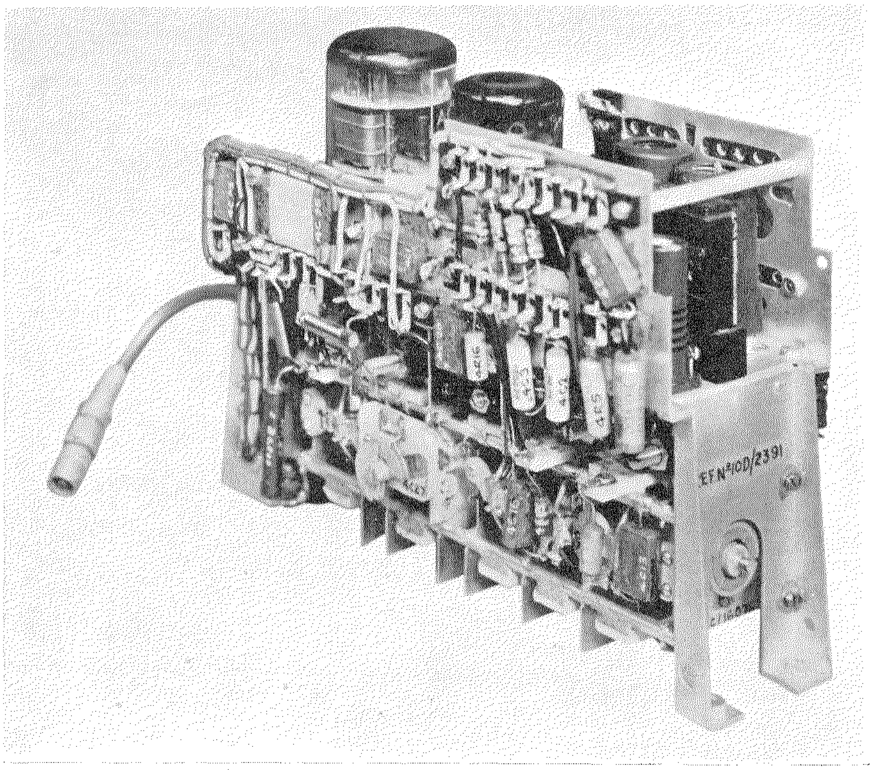

Fig. 11—Transmitter. The coaxial cable goes to the antenna change-over relay. The 5-gang tuning capacitor is at the bottom.

modulated up to 100 per cent. Harmonic output will not exceed 15 per cent for modulation of 80 per cent. The noise level is at least 40 decibels below the level corresponding to 100-per-cent modulation. The audio-frequency response, relative to values obtained at 1000 cycles, will be between −4 and −10 decibels at 300 cycles and between zero and −4 decibels at 3000 cycles.

The transmitter may be seen in Fig. 11 and consists of an auxiliary crystal oscillator at half the intermediate frequency (or 4.86 megacycles), a balanced radio-frequency modu-

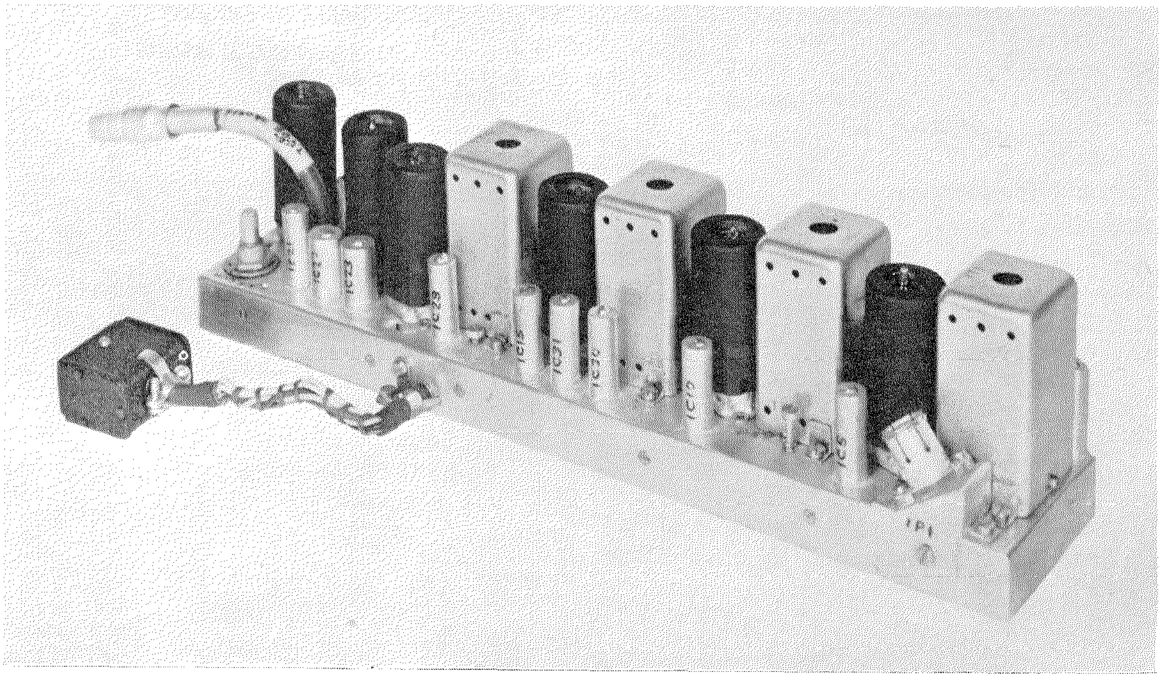

Fig. 12—Intermediate-frequency amplifier. The inter-stage coils are in the rectangular cans and the valves are in the cylindrical shields.

lator, doubler, amplifier, and power-output stages, the latter using a double-tetrode valve. Mixing of the input from the second trebler stage of the receiver (at 9 times the fundamental crystal frequency) with the output from the auxiliary crystal oscillator is effected in the balanced modulator.

Output from the doubler stage is band-pass coupled to the output amplifier where it is modulated by the audio-frequency signals received from the modulator unit.

In physical construction, the transmitter is generally similar to the receiver, being built in the same unit-type construction around a similar 5-gang tuning capacitor together with all valves, coils, screens, trimmer capacitors, test plugs, etc., the capacitor again being connected to the channel-selection mechanism by a flexible coupling.

3.4 INTERMEDIATE-FREQUENCY AMPLIFIER

Three stages of intermediate-frequency amplification at 9.72 megacycles are incorporated in this unit, using high-slope variable-mu pentodes, and these are followed by a diode detector with which are associated a squelch stage and an automatic-volume-control amplifier. The squelch circuit feeds an audio-frequency amplifier in the modulator unit for the telephone output from the equipment, whilst the automatic-volume-control amplifier provides a voltage that is applied to the screens of the various intermediate-frequency stages. A portion of this voltage is tapped off and used to control the radio-frequency-amplifier stage in the receiver unit.

The interstage coils are inductively tuned and are provided with convenient trimming devices.

The intermediate-frequency-amplifier chassis, shown in Fig. 12, is of silver-plated brass in the form of a shallow inverted box, with valves, screened-coil units, decoupling capacitors, and coaxial connectors mounted on top. Other components mount below.

3.5 MODULATOR

This consists of a microphone-amplifier stage, a telephone-amplifier stage, and a push-pull amplifier supplying modulation to the transmitter. Negative feedback is employed between the telephone and microphone amplifiers to avoid distortion and to maintain good regulation.

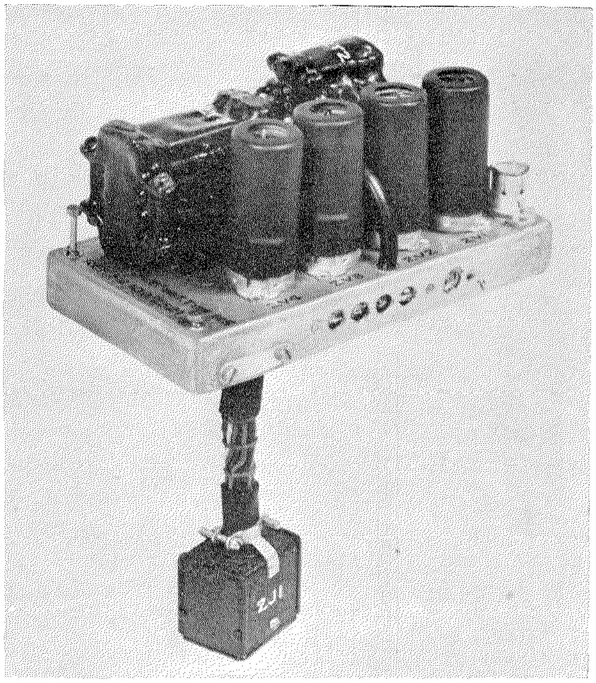

Fig. 13—Modulator. The modulation, driver, and microphone transformers are at the rear.

The chassis is of inverted-box construction in aluminium alloy, with valves and transformers mounted above the deck and other components below. It may be seen in Fig. 13.

3.6 POWER SUPPLY

Plate power at 250 volts and grid bias at 50 volts are generated by a rotary transformer mounted on a jack-in-type aluminium-alloy chassis, which also forms a screened box for fuses and input and output electrical filters. The air blower incorporated in the machine consists of a bakelite centrifugal impeller shrouded by an aluminium spinning conforming to the diameter of the machine. A rubber gasket is fitted to engage with the air inlet of the main-chassis dust cover when the power-supply unit is in position.

The mechanical drive unit at the other end of the rotary transformer was also designed to conform to the machine diameter to make a compact overall assembly.

3.7 CHANNEL-SELECTION MECHANISM

The channel-selection mechanism is constructed as a self-contained unit fitted externally to the front panel of the main chassis. It is connected mechanically to the receiver and transmitter tuning-capacitor spindles and to the output shaft of the drive unit on the rotary transformer by double flexible couplings to allow for misalignment of shafts.

Incorporated in the unit, which may be seen in Fig. 14, are a mounting for the four crystals and a bank of four slide switches, which connect them in the circuit as required.

The essential parts of the mechanism are two setting spindles for the receiver and transmitter capacitor drives and four spring-returned slides, which rotate the setting spindles through the medium of cam levers to positions corresponding to the channel selected. Operation of the slides brings the appropriate crystal into the circuit via its slide switch at the same time as the capacitor spindles are positioned. A camshaft, driven from the drive-unit output shaft, operates the slides in sequence until the correct slide has been brought into use, when the camshaft is automatically tripped by the magnet in the drive unit. Positioning of the camshaft is controlled by two switches driven by the camshaft itself. The first is of a conventional rotary type and is used only to discriminate between the electrical markings transmitted from the remote control box for the four channels. The second switch comprises snap-action contact springs operated by four cam-notches in a large-diameter disc on the camshaft. These notches correspond to the stopping positions of the camshaft for correct operation of the four slides and control the drive-unit magnet directly.

Another set of springs actuated by the same four cam-notches enables the cam-shaft to be driven to a "free" position in which all slides are released but can be operated manually for setting and resetting purposes.

Further switches are provided on the mechanism to enable the settings of receiver and transmitter to be checked by meters.

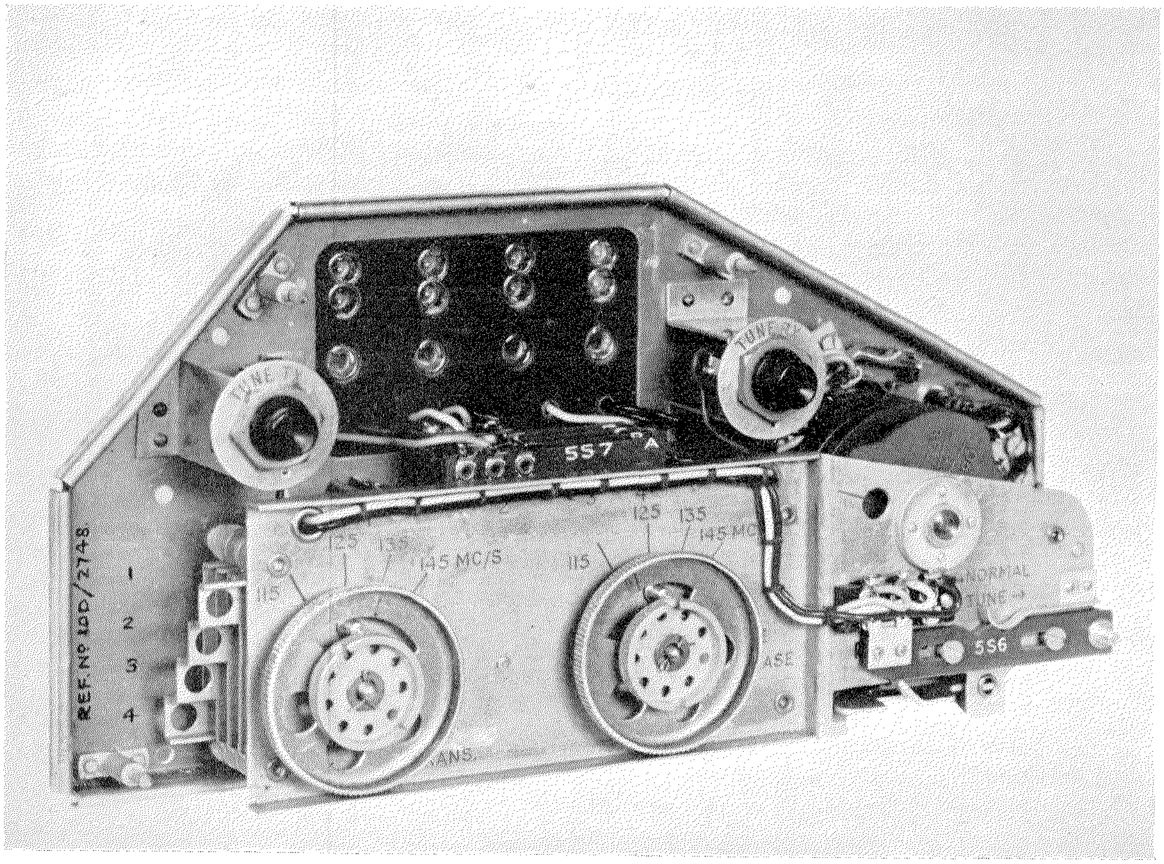

Fig. 14—Channel-selection mechanism. The frequency-setting spindles are at the bottom centre, and above them are the mountings for the 4 crystals.

The mechanism incorporates a compensating cam device that smooths out to a large extent the fluctuating torque imposed on the camshaft by the spring-loaded operating slides.

Although the camshaft stops within approximately 1.5 degrees of its correct position, the need for special accuracy at this point is obviated in the design. Furthermore, the use of spring loading and vee or diamond bearings eliminates setting inaccuracies, which would otherwise occur due to manufacturing tolerances in the various parts, and enables the ganged tuning capacitors to be positioned to an accuracy of ±0.125 degree.

The mechanism is arranged to permit the unlocking, resetting, and relocking of the setting-spindle positioning for any channel of either receiver or transmitter without disturbing the existing settings for other channels.

Duralumin is used for lightness, rigidity, and wearing properties in the construction of the slides and mounting arrangements.

A detachable aluminium waterproof cover fits over the complete mechanism, with a built-in Perspex window over the crystal mounting for inspection purposes. When the cover has been removed for resetting or similar reasons, any switches inadvertently left in the wrong position are either automatically restored to normal in refitting the cover, or are arranged to foul the cover until restored by hand.

3.8 Remote Control

The remote control unit used with the equipment is a small bracket-mounted aluminium-alloy box fitted with a plug connection for external cabling and a single switch for turning the set on and off and for channel selection.

This unit was purposely kept as simple and small as possible to reduce operating procedure to a minimum and to facilitate installation of the control unit in the congested cockpits of small aircraft.

3.9 Weights and Dimensions

Table I gives the weights and dimensions of the component units of the A.R.I. 5272 set.

TABLE I
Weights and Dimensions

Unit	Weight		Dimensions (Height × Width × Depth)	
	Pounds-Ounces	Kilogrammes	Inches	Centimetres
Receiver	1–13	0.8	5.1×2.0×6.9	13×5×18
Transmitter	2–10	1.2	6.0×3.25×8.0	15×8×20
Intermediate-Frequency Amplifier	1–12	0.8	2.9×2.9×10.75	7×7×27
Modulator	1–12	0.8	2.8×3.1×5.7	7×8×14
Chassis, Including Channel-Change Mechanism, Carbon-Pile Regulator and Rotary Transformer	13– 9	6.2	7.9×9.0×16.25	20×23×41
Control Box	0– 8	0.2	3.5×2.2×2.6	9×6×7
Total, Less Cables	22– 0	10.0		

Survey of the Telephone Transmission-Rating Problem*

By L. C. POCOCK

Standard Telephones and Cables, Limited, London, England

THE QUALITY of service provided by a telephone concern is an essential factor in its financial economy. In this relation, it is necessarily the overall quality as experienced by the subscribers. Division into different aspects such as traffic quality and transmission quality is arbitrary; each may be separately measured, but for economic balanced planning there is need of a combined measure of overall quality. The paper deals only with transmission quality. Planning has, in the past, been in terms of circuit attenuation and instrument efficiency, with the object of ensuring an adequate level of received speech, but this conception is no longer adequate or economical.

The direct observation of the number of times a subscriber finds it necessary, in a unit time-interval, to repeat a word or a sentence is the best known indication of transmission quality in service; a complete rating system based upon such observations† and other considerations has been put into operation by the American Telephone and Telegraph Company.

In Europe, the repetition-rate criterion of service quality has been accepted, but application of the practice presents difficulties. Two plans are being investigated: (*a*) attempts are being made to find, under certain restricted conditions, a correlation between repetition rate in service and laboratory articulation tests; and (*b*) a provisional system of rating based on articulation tests (without reference to repetition rate) is being studied.

Both plans are described, and possible correlation procedures are discussed. Reference is made to the economy aspect of rating on the broader basis of transmission quality, but the subject is too large to treat adequately in the present paper.

Brief mention is made of foreign publications.

Appendices include detailed descriptions of the experimental procedure to establish transmission-service ratings.

On account of the fluid condition of the subject of the paper, standardized terminology is not in existence and therefore definitions of the terms used are given as a final appendix.

• • •

1. Introduction

The fundamental economy of industrial concerns comprises cost of production, quantity of production, and selling price. The business of providing telephone service rests upon similar fundamentals, which may be more appropriately described as:

A. The cost of providing service, expressed as total annual charges.

B. The extent to which the service is utilized, evaluated in call units.

C. The charge made to the user per call unit.

These three interdependent elements may restrict the activity of the engineer, but he can receive no positive help from them until there is added a factor directly related to the accounts and susceptible to control by engineering design.

This fourth factor is the quality of the service given to the public; it is the touch-stone of the engineer's contribution to the telephone business and of the business itself as an alloy of finance and engineering.

Fig. 1 illustrates the kind of relation that must exist between the four quantities mentioned. The fundamental importance of such relations is obvious: not only does the measurement of service quality become important, but essentially the quality must be correctly measured, i.e., in terms correlating with the subscriber's experience, because it is the subscriber's satisfaction

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* Reprinted from *Journal of the Institution of Electrical Engineers*, v. 95, Part III, pp. 253–265; July, 1948.

† One of the difficulties of obtaining repetition-rate data is the obligation not to infringe the privacy of subscribers' calls. The existence of two large bodies of employees at the Bell Laboratories and in the telephone company's building, with considerable tie-line traffic between them, enables the American Telephone and Telegraph Company to make observations under substantially service conditions without tapping subscribers' lines.

with the service that influences the extent to which he will use it and pay for it.

1.1 QUALITY OF SERVICE

Since it is the purpose of a telephone business to provide a satisfactory quality of service at the lowest cost, there arises at once the searching question, What is a satisfactory service?

At any given epoch, the quality of the service may be improved at a certain increase in cost, and it is common experience[1] that the use of the service increases when the service is improved, so that increased revenue partly offsetting the cost may sometimes be obtained without any increase in charge.

The process of improving the service may be continued with or without aleration in charge until further improvement would result in reduced use on account of the high charges necessary. The quality of service is then theoretically the best economically obtainable; other considerations might, however, dictate a policy of a rather lower standard at lower cost.

According to the principles of cost equilibrium, the annual charges will be least for a given quality of service when the increase in annual charge, due to making a given small improvement by making changes in any one of the parts into which the system may be arbitrarily divided, is the same whichever part is selected for amelioration. From this it follows that the contribution of all parts of the system to the overall service result is determined by economic principles, which enable a choice to be made between the various technical alternatives that might be proposed to provide the grade of service required. It follows that it is necessary to provide means to measure the quality of service and to evaluate the contribution of each part to the quality of service of the whole system.

It must be remembered that service quality thus far discussed is measured by the subscribers' appreciation, which is determined not only by transmission performance but also, amongst other things, by speed, availability, and reliability of service, which may be arbitrarily called "switching quality."

If, therefore, a necessary division is made between transmission quality and switching quality, there is a need to create a new unit of "overall service quality" which will be a function of these two components, combining them in suitable proportions. For example, the units may be so chosen that a unit change in transmission quality in service is equal to a unit change of switching quality, if both produce the same effect on the use of the telephone. It is evident that, if the system is to be so planned that there is economic balance between the expenditures on switching and on transmission with maximum overall satisfaction to the subscriber, some such relation is required to equate switching and transmission quality.

2. *Quality of Transmission*

The paper is concerned with quality of transmission, which is the aspect of overall quality of service perceived when the quality of switching service is disregarded. Transmission quality in service is here called "transmission service" and its numerical values are called "transmission-service ratings." It will be necessary to consider how it is to be assessed (rated) and how the overall rating may be broken down to determine the contribution of every part of the circuit to the total transmission service. It will facilitate the the study, however, to recognize at once that it is more difficult to give good quality over a long circuit than over a short one, and in general that losses in quality or quality impairments relative to some standard are contributed by each part of a circuit. The study of transmission service is

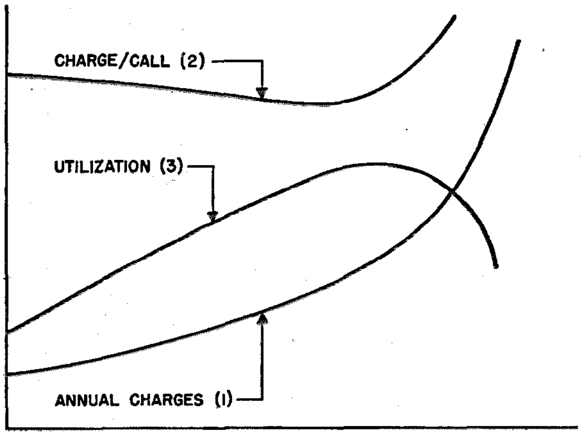

Fig. 1—Fundamental relations. (1) = (2) × (3).

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[1] Bibliographical references will be found on page 272.

therefore to be conducted by the study of the complementary idea of quality impairment in service, and it becomes a further object of study to express quality impairments in such a way that the contributions of the parts of a circuit may be added together or combined in some simple way to give the quality impairment of the whole circuit.

In the past, transmission circuits have been planned to have predetermined overall attenuation with only qualitative attention given to the amount of speech distortion and other factors. The planning of circuits on this basis was a technical matter and the technicians concerned could be sure that the circuits planned would be satisfactory if the subscribers behaved in a prescribed manner. Such a system is unsuited to modern requirements. First, it is not economical to design speech circuits merely to give sufficient loudness because it is not to the advantage of the subscriber and does not tend consistently towards better transmission quality, nor does it provide a required grade of transmission at the lowest cost. Secondly, it is just as true for monopoly services or public utilities as for competitive industry that sales can be increased by studying the consumers' habits and requirements and adapting the service or commodity offered to suit them. Finally, when the subscribers' behaviour is studied, it is found that it is an important factor in the performance of a given physical circuit, and that the average behaviour can be shown to depend at least in part upon some of the characteristics of the circuit. W. H. Martin[1] has discussed in detail the adjustments of behaviour that occur in direct and telephone conversations. The differences between the two cases arise mainly from the differences between the circumstances (e.g., in respect of room noise) of the two persons talking by telephone, as opposed to the close similarity of circumstances for two persons in the same room, and from the ignorance of a speaker on the telephone of the surroundings of his listener.

Apart from these facts, which are external to the telephone, the artificial sidetone introduced by the telephone and the artificial channel of communication itself, which may be unequal in efficiency in the two directions, produce reactions upon the manner of speaking. Some of these reactions may adversely affect transmission quality and it is therefore desirable to eliminate as far as possible those causes which give rise to unsuitable reactions, and to utilize as fully as possible favorable reactions.

To sum up, transmission quality in service is the transmission data that is relevant to the economics of telephony; it is the resultant of objective circuit characteristics and subscribers' actions. Subscribers' actions are, on the average, largely determined by the circuit characteristics and therefore need not be explicitly known if transmission quality is directly measured in service; the results thus obtained may be expressed as a function of circuit characteristics, the subscribers' reactions to the circuit being implicitly included by the measurements. Knowledge of the subscribers' contribution to transmission performance is, however, useful in analysing the results of service observations, in assisting design to eliminate unfavourable subscriber reactions, and in leading to better simulation of direct speech conditions in telephony.

The study of transmission performance in the broad sense of appreciating the quality of transmission obtained in service, leads to the establishment of transmission-service ratings as numerical values for the impairments contributed by parts of a circuit.

3. Circuit Design on the Basis of Transmission Service

3.1 DATA REQUIRED FOR CIRCUIT DESIGN

The following are the requirements for solving the problem of designing circuits on the basis of their transmission service:

A. The complete technical data of each circuit must be known, i.e., attenuation, distortion, noise, sidetone, etc., must be subjectively or objectively measured. In this way, each circuit can be fully specified in terms of a certain number of parameters, which are preferably chosen and defined in such a way that they are independent of each other.

B. A measurable quantity must be found that is uniquely related to the transmission quality impairment as experienced by subscribers in service.

C. Using B as a measure and varying in turn each of the parameters mentioned in A, relations must be found between the transmission quality impairment and different values of each circuit parameter.

D. Means must exist for combining into one total the transmission quality impairments attributed to the values of the several parameters. The total result thus obtained must agree with the direct assessment of transmission quality impairment when a circuit having several parameters different from the normal is used in service and measured in terms of the quantity referred to in *B*.

E. For the purpose of practical engineering planning, it must be possible to assign transmission impairments to the separate links of a circuit that may be joined together, in such fashion that the total impairment for a circuit built up from two or more such links can be determined from the impairments assigned to each link.

The above requirements will now be considered in turn.

3.2 Circuit Parameters

The circuit parameters required for the objective description of a circuit have been chosen by the American Telephone and Telegraph Company, as follows:

Attenuation of the circuit.

Attenuation of the sidetone path for speech at the transmitting end.

Attenuation of the sidetone path for noise at the listening end.

Efficiency of the receiving end for line noise.

Circuit distortion.

with the additional quantities:

Amount of line noise.

Amount of room noise.

It does not seem possible to choose a better set of quantities to describe the characteristics of a circuit; all the quantities can be so defined as to be independent and each is capable of subjective or objective measurement. It should be noticed, however, that from the transmission point of view practically all these quantities extend over the range of transmitted frequency, and when they are expressed by a single number, some convention or particular application of the measurement is implied. The tendency is to develop as far as possible objective, rather than subjective, measurements using frequency weightings such that the results agree with subjective measurements. Such methods save time and are more easily repeated.

3.3 Measurable Quantities that Concern a Subscriber

It has been shown that the aspect of transmission quality that is significant can be observed only in service conditions; it is therefore necessary to look for measurable quantities in which the subscriber is concerned. The choice is limited, and the following is believed to be a complete list:

A. Repetition rate. On a monitored circuit, observations may be made of the average frequency of occurrence of requests for words or sentences to be repeated.

B. Unfavourable comments. On a monitored circuit, observations may be made of the average frequency of occurrence of remarks between subscribers criticizing the quality of the transmission.

C. Transmitter output. Observations can be made at the local exchange of the transmission level that, when corrected for local-line loss, depends on the subscriber's voice level and the closeness with which he speaks to the transmitter.

D. Average duration of call. A small amount of evidence has been obtained showing that when circuit conditions are changed the average duration of calls changes (see Section 10).[3]

Of these four manifestations of the subscriber, only the first two seem to be simply and unequivocally related to the subscriber's experience of difficulty or dissatisfaction with the circuit he is using. Observations of transmitter output may provide useful additional data and, in some cases, transmitter output reflects the effort the subscriber has to make, but it is also known to depend on unconscious adjustment of voice level depending upon the amount of sidetone and the amount of room noise. There is no clear evidence at present that subscribers respond to circuit defects due to distortion by speaking more loudly or closer to the transmitter.

Observations of average duration of call can be very easily made automatically; at present insufficient data exist to show whether there is a clear relation between this quantity and circuit parameters.

It has been agreed, internationally,[4] that "repetition rate" indicates the subscriber's degree of dissatisfaction with a circuit; "unfavourable comments" appear to be far less satisfactory because they begin to give an indication only when the circuit is already bad.

Repetition rate is not an absolute measure, but it can be used as a criterion of equality by adopting the rule that circuits having the same repetition rate for a given group of subscribers are equal in performance from the subscribers' point of view.

Repetition rate being adopted of necessity and by agreement as the criterion of equal circuit performance, a sufficient number of subscriber stations must be put under monitored supervision, and repetition-rate counts must be made while the different circuit parameters are successively varied from a typical reference value over a sufficiently wide range, one at a time.

From the results, curves can be drawn showing repetition rate against variations of each parameter from its reference value. All the curves will show the same repetition rate for the reference value of each parameter and therefore at any other value of repetition rate the corresponding values of parameters are equivalent in the sense that they produce the same change in repetition rate. It is convenient by this process to relate parameter values to changes of distortionless attenuation, which are equivalent in the sense described; these attenuation values in decibels are then the transmission impairments brought about by the variation from reference value of the parameters from which they are derived. By this procedure, all impairments are expressed in decibels and repetition rate is used only as a criterion of equality (Section 14.1).

3.4 ASSESSMENT OF THE QUALITY OF A COMPLETE CIRCUIT

When the impairments expressed in decibels are known for all practical values of the circuit parameters, it is a natural step to add together the appropriate impairments for any given combination of circuit parameters. Such addition cannot be pushed too far, but it may be kept within reasonable limits by choosing, for reference, values in the middle of the field of practical variation.

The breaking down of the total impairments of a complete circuit into impairments assigned to each link of the circuit so that the impairments of any such links can subsequently be added to give the total impairment of a new circuit, requires the establishment of a reference circuit. Mention has been made above of reference conditions; the adoption of one particular fully specified circuit as a reference circuit provides a complete standard of comparison. The results of measurements of sidetone, line noise, etc., are expressed relative to the values measured in the selected reference circuit; similarly the impairment of, for example, a complete transmitting system (telephone set, local line, and cord circuit, considered in the talking direction only) is observed relative to the corresponding part of the reference circuit. In effect, any circuit link to which an impairment is to be assigned is substituted in the reference circuit for the corresponding link, and the change in circuit performance expressed as an equivalent change in reference-circuit attenuation is the impairment of the substitute part.

It is an essential requirement that the impairments assigned to the links of a circuit should add up to the impairment of the whole circuit as determined by direct comparison with the reference circuit.

4. *Transmission-Service Rating (U.S.A.)*

The system of transmission-service rating that has been outlined in Section 3 is the system devised and put into operation under the name of "effective transmission rating" by the American Telephone and Telegraph Company. A more detailed account has been published by F. W. McKown and J. W. Emling.[2]

On the subject of repetition rate, W. H. Martin[1] wrote in 1931:

> "Different degrees of success in carrying on telephone conversation may be taken as being indicated by the number of failures to understand the ideas transmitted over the telephone and by the amount of effort required on the part of the users to impart and receive these ideas. . . . The repetitions required in a conversation can be noted but a determination of the effort factor presents difficulties."

Martin also advances the following arguments:

A. Circuits which show equal repetition rates are likely to call for equal effort.

B. When the effort appears to the subscriber to be excessive, his opinion of the service will be adversely affected and the "adverse comments" on the transmission, which he will make, can be recorded by the observers and used to supplement the repetition rate in arriving at a better picture of the service.

The subject of effective transmission rating was developed in further detail by F. W. McKown and J. W. Emling[4] in 1933, where the authors state:

> "Properly designed, a telephone system should minimize to the degree consistent with costs its inherent differences from direct conversation, and make it easy for the ordinary user to get, without undue effort, results which are satisfactory to him in comparison with direct conversation."

Attention is directed to these two quotations because in concentrating on the repetition-rate aspect of effective transmission the subscriber's effort is apt to be overlooked and forgotten. It is desirable that data should be collected on this subject in so far as it is possible to do so; to give a simple example, it should be possible to determine whether the subscriber's voice effort on circuits that have such high attenuation as to require loud speaking is related to the repetition rate and whether there is a tendency to speak equally loudly on circuits giving a similar repetition rate when the impairment is mainly due to distortion.

5. *Transmission-Service Rating (Countries Other Than the U.S.A.)*

It was only after very considerable discussion that the Comité Consultatif International Téléphonique placed on record[4] in 1938 the inescapable conclusion that repetition-rate observations afford the best known measure of transmission performance.

The spread of the practice of transmission-service rating to other countries from the U.S.A. has been slow. The reasons for such slow development are associated with the geographical and technical magnitude of the task; the extensive series of fundamental observations of repetition rate made in America is of great value as research in fashioning a new tool and indicating the bearing of circuit conditions on transmission performance. The actual results are, however, empirical in character and applicable only to American sets, plant, and conditions, so that they do not provide data directly applicable to other countries. Therefore, in every country a similar extensive programme of fundamental observations would have to be carried out independently.

This would be for each country a big programme of work involving the installation, supervision, and maintenance of a large number of sets with special controllable features, and a very long series of repetition observations. In some countries, more than one type of set is in use, and this would increase proportionally the programme of work. Such an undertaking might be justified by the end in view. The mere accumulation of data in each country, however, would be simple compared with the task of co-ordinating the data, which would be possible only by installing identical transmission systems in every country and making service observations on these.

On quite another plane, the large number of organizations independently engaged in development work have a natural desire to retain within the domain of the laboratory the testing procedure and the criteria that determine what is or is not a significant impairment in transmission. Although two leading countries carried out considerable experimental work on the observation of repetition rate, it was apparent that the full programme of work needed was beyond the present capacity of the majority of administrations, and alternative plans were studied for achieving the same result by simpler methods. The procedure adopted was the outcome of an offer of co-operation from the American Telephone and Telegraph Company.

It was known that laboratory articulation tests, which had long been used for the comparative study of transmission instruments, did not correlate satisfactorily with repetition-rate observations; obvious reasons why correlation could not be expected were the artificial restrictions placed on articulation tests where fixed values of voice intensity and talking distance must be adopted in order to give meaning to results obtained by a small band of testers. In service, on the other hand, very large numbers of speakers and listeners are involved; the voice level and talking distance vary over wide ranges and are statistically stable, though the mean values vary from one set of circuit conditions to another.

It was suggested, therefore, that correlation might be found if articulation scores were compared with repetition rates, the received loudness level in the articulation tests being the same as the average received loudness level in service when repetition rates were observed.

The American Telephone and Telegraph Company placed at the disposal of the Comité Consultatif International Téléphonique some of the special sets with controllable features that had been used for the fundamental series of repetition-rate observations made on calls by employees over tie lines between the Bell Telephone Laboratories and the American Telephone and Telegraph Company building in New York; they undertook to supply full information regarding room noise, line noise, and average voice-level as determined by volume-indicator readings. With this information, it was possible for the Comité Consultatif International Téléphonique laboratory to make articulation tests under conditions closely resembling the average conditions under which repetition rates had been observed.

This programme has been substantially completed, but certain details require further study and the results cannot be made public until they have been carefully analysed; the writer is of the opinion that sufficient evidence of correlation will be found to justify further study.

Should a satisfactory measure of correlation be found between repetition rate and articulation at equal received levels, considerably more work will be required before transmission-service rating based on articulation results and received-level data can be equated to effective rating (repetition-rate basis). It will be necessary to determine whether the particular correlations found for the American sets are valid for different types of European sets and conditions. For this purpose, repetition-rate data will have to be determined in two or three European countries. Further, it will be necessary to establish a considerably higher standard of agreement between articulation results obtained on the same circuits in different laboratories than has hitherto been attained.

6. Possible Relations Between Repetition Rates and Articulation Tests

There seems no reason to doubt that articulation tests and other similar closely allied laboratory tests, carried out under the usual recognized conditions, are a useful measure of the inherent capability of the circuit tested to transmit speech. It is because only a comparatively small number of individuals can be used in such laboratory tests that the wide range of influence on service results exercised by actual subscribers must be excluded by substituting controlled talking and listening conditions.

In seeking for a relation between repetition rate and articulation percentage, the problem is therefore to know and allow for the subscribers' behaviour. Suppose that a particular circuit is considered which is used by a large number of subscribers: some will talk more quietly than others or at a greater distance from the transmitter; some will place the receiver less firmly on the ear, or they may have appreciable hearing deficiency. The same circuit will therefore be used with varying degrees of success by different persons. It is not difficult to imagine that some users will find the circuit satisfactory with low repetition rate, while others will experience a substantially higher repetition rate; the repetition rate recorded in service observations will be a mean value taken over a range of added "personal" impairments.

The personal impairment here referred to is the sum of the subjective and objective losses for which a particular subscriber is responsible by his manner of speaking and listening, and it is reckoned relative to the mean talking intensity, talking distance, and hearing acuity of all the subscribers.

When the same circuit is set up for articulation tests in the laboratory, a standardized talking intensity and talking distance are adopted. In order that the laboratory result may be obtained at the same received level as the speech on which repetition observations were made, the attenuation used in the laboratory test must generally be varied from the real circuit attenuation to allow for the difference between (*a*) the standardized talking level and distance, and (*b*) the average talking level and distance in service; i.e., an allowance must be made for the difference in the transmitter outputs in the articulation tests and in service, the latter being obtainable from service observations and dependent upon the circuit parameters and especially the sidetone.

From various points of view it has been proposed that articulation scores, obtained over substantially longer lines than are used in service, should be considered as significant for assessing transmission performance. Such a procedure leads to correlation being sought with repetition rate, not at the same received level, but with articulation observed at a received level lower by a constant number of decibels; this may be called "correlation with added loss."

This idea may be justified by the consideration that the members of an articulation team are usually selected individuals with good hearing, whereas surveys have shown that a certain deficiency of hearing is widely distributed throughout the telephone population. It is therefore reasonable to add attenuation in the articulation tests to represent the average deficiency of hearing of the population, or, if articulation has been studied as a function of circuit attenuation, correlation may be sought between repetition rate and articulation at a received level lower than that observed in service by a constant number of decibels.

Surveys have indicated[6] a standard deviation of hearing loss of at least 8 decibels, and it is suggested that the "added loss" required for correlation with repetition rate should be at least twice the standard deviation.

It must be mentioned, however, that insertion of added loss in the line is a very imperfect counterpart of hearing loss, since it attenuates the received speech without attenuating the room noise heard by the listener; more properly the added loss should be introduced in series with the telephone receiver, or the room noise should be appropriately reduced in intensity.

While it is undesirable to introduce unnecessary complications, the usefulness of establishing correlation between repetition rate and articulation is so great and the differences between service and laboratory conditions are so considerable that it may be worth while to study a composite function of articulation which takes these differences more explicitly into consideration.

If a series of articulation tests is made on a given circuit producing results illustrated by curve (1), Fig. 2(*a*), and then further tests are made under conditions of greater noise, the results will be of the kind shown by curve (2). A further series of tests on the circuit as first tested but with a low-pass filter inserted will give results of the kind shown by curve (3).

If the empirical law $N = 50 \log_{10} R/R_0$, which has been quoted as valid for these types of impairment, holds over a sufficient range, the corresponding repetition-rate curves will present the appearance of Fig. 2(*b*). In these circumstances, consistent correlation between repetition rate and articulation, either direct or with added loss, cannot be obtained over a range of attenuation values.

Since part or most of the subscribers' contribution to overall transmission performance is the introduction of a range of personal impairments over which repetition rate must be averaged, it is more likely that the desired correlation will be found if articulation is similarly averaged over a range of values of attenuation, especially when (as in Fig. 2) different kinds of impairment are considered. This leads to comparison of repetition rate with the weighted average of articulation taken over the range of personal impairments. The combination of this process with "added loss" is illustrated in Fig. 3, where the dotted curve representing the distribution of personal impairments may be taken as Gaussian with a standard deviation of about 10 decibels.*

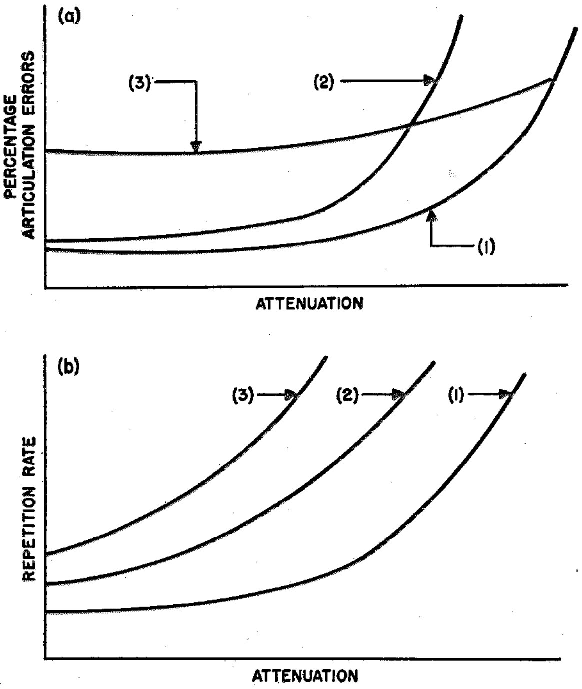

Fig. 2—Variation with attenuation of (*a*) articulation errors, (*b*) repetition rate on three different circuits.

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* This value is obtained from a standard deviation of 6 decibels for observations of transmitter output (Appendix 14.5) combined with a standard deviation of 8 decibels already quoted for hearing loss.

The weighted value of articulation errors $\bar{A}$ is the value to be used for attempted correlation with repetition rate observed at the average received level x_1.

The special significance of this proposal is that it takes into consideration the transmission performance for all the subscribers, as it should do in accordance with the economic principles that have been stated, and is not confined to observations on lines of limiting performance.

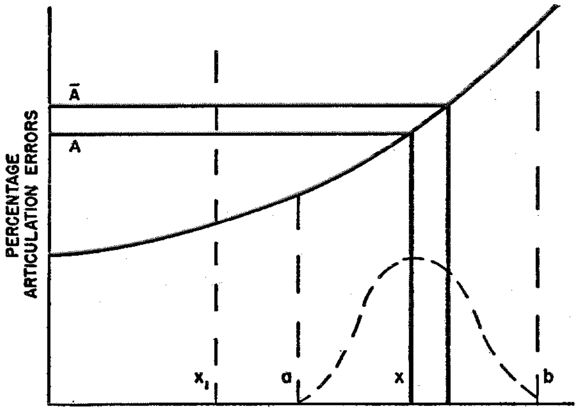

Fig. 3—Illustration of weighted average articulation errors.
$x - x_1$ = **Added loss.**
$b - a$ = **Range of personal impairments.**

There is still one respect in which articulation tests fail to represent service conditions. It is well established that there is a considerable loss of hearing in the upper frequencies with advancing age, and therefore articulation-testing teams should be selected to have age distribution representative of the telephone public, or preferably a range of networks should be used to reproduce the appropriate hearing losses at upper frequencies.

7. *Transmission-Service Rating by Articulation Tests*

In view of the long period that must elapse before European transmission-service ratings uniform with effective ratings can be applied, and the urgent need to introduce an improved system of rating at an early date because of the large amount of telephone development and reconstruction work resulting from the war, the Comité Consultatif International Téléphonique decided[7] in 1946 to attempt the development of a provisional system of transmission-service rating based upon articulation and received level. The proposal is an anticipation of the results of the study of correlation between repetition rate and articulation results, which has been described, because repetition rate does not enter into consideration, and no proof is required that the new ratings are intimately in accordance with the Comité Consultatif International Téléphonique fundamental declaration of the validity of repetition rate for measuring transmission quality in service. The introduction of new ratings on the proposed lines is justified by the consideration that such ratings must inevitably be nearer to the desired results than the volume ratings hitherto used.

The new ratings must, however, fulfil the practical requirement that impairments thus determined for parts of a circuit can be added together and give the same total as can be determined when the circuit is rated as a whole by the same method.

If a successful correlation can be achieved between repetition rates and articulation scores at related levels using one of the correlation techniques that has been described, the dictum "circuits with equal repetition rates are equal in effective transmission," will become "circuits which have equal articulation (reckoned according to the requirements of the successful correlation method) differ in transmission service rating only by the amount of any difference between the transmitter output in service and the transmitter output in the articulation tests."

In the present circumstances, it is not permissible to assume that such extensive correlation exists over a range of articulation values. A simpler and seemingly safer concept has accordingly been set up.

Instead of endeavouring to rate a circuit by its average performance for all subscribers, the proposal is to rate the circuit by its performance under the worst conditions to be encountered in reasonable service. This is equivalent to inspecting a manufactured product to an individual minimum requirement, in contrast to using quality control as a means of maintaining a given standard of production.

It has been seen in Section 6 that the nominal attenuation of any circuit in service is supplemented by a wide range of positive or negative personal impairment. The increase in attenuation due to the effects of noise, hearing deficiency, subscriber's voice level, and speaking distance may amount to 35 decibels or more in extreme cases; making conservative assumptions, 25 decibels may be taken as a limit exceeded by $2\frac{1}{2}$ per cent of subscribers. If, therefore, under the controlled conditions of laboratory articulation tests, additional attenuation, which may be of the order of 25 decibels, is added to the circuit, articulation measurements should indicate the worst results to be experienced among $97\frac{1}{2}$ per cent of subscribers. It remains to decide on the grade of service that should be encountered at this limiting condition and specify it as a fixed articulation percentage X. Then, any circuit that gives the articulation X is, after removal of an amount of attenuation equal to the added loss, a limit circuit to which an articulation rating of a convenient constant number of decibels can be assigned.

This rather cumbersome statement can be made clearer by an example. A commercial transmitting system can be defined as a battery feed circuit, a local line, and a telephone set considered in the transmitting direction; a similar system in the receiving sense is a commercial receiving system. If several different transmitting and receiving systems are connected in pairs through distortionless attenuation, curves of articulation against attenuation can be plotted as in Fig. 4. The transmitting and receiving systems represented by curves (*a*) and (*b*) may be connected through N_a decibels and N_b decibels, respectively, before reaching the limiting articulation X. Therefore the transmitting and receiving systems of (*b*) are together $(N_a - N_b)$ decibels worse than the transmitting and receiving systems of (*a*).

It must be remembered that all circuits giving articulation X are to be regarded, after removal of the added loss, as having equal articulation equivalents, say N_0 decibels; therefore, if $(T + R)_a$ is written for the articulation equivalent of the (*a*) transmitting and receiving systems, with a similar notation for the (*b*) systems, it is possible to write

$$(T + R)_a + (N_a - L) = N_0,$$
$$(T + R)_b + (N_b - L) = N_0,$$

where L is the amount of the added loss.

The values of $(T + R)_a$, $(T + R)_b$ obtained from these equations are called articulation reference equivalents because of the method used to determine them; their magnitudes depend upon volume loss as well as distortion. The volume loss is that of the circuit under the controlled conditions of articulation testing and therefore the articulation reference equivalents have to be adjusted by an amount equal to the difference between the transmitter output in the articulation tests and in service. When this has been done, the adjusted articulation reference equivalents become transmission-service ratings.

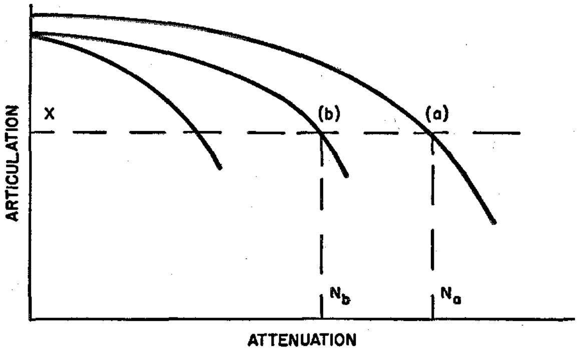

Fig. 4—Articulation equivalent. The circuit represented by (*b*) has an articulation equivalent $N_a - N_b$ relative to the circuit represented by (*a*).

The fixed quantities involved in this plan are a limiting value of articulation X, a value of added loss L, and a convenient number N_0 for the limiting rating. It has already been indicated how L may be arrived at, and N_0 should be selected so that the numerical values representing new ratings will not differ more than is necessary from the old volume ratings for the same circuits.

The value to be assigned to X should be found by articulation tests on commercial types of circuit when L has been fixed. It is proposed, however, that X should be an intermediate quantity and not the formal specification of equal articulation equivalents, because, as is well known, articulation values are not absolute and not exactly reproducible. Instead of specifying X, it will be better to specify an exact physical circuit (articulation reference circuit) so adjusted as to give approximately the required articulation; the conditions of equality as illustrated in Fig. 4 will then be deemed to be satisfied when articulation on a tested circuit is equal to the articulation on the specified reference circuit.

8. *Articulation Reference System*

There has been in Paris for some 20 years a transmission system of known and specified performance, which has been the central standard for the volume ratings hitherto in use.[8] It is convenient to adapt this system for use as the European articulation reference system.

The modifications at present contemplated are the introduction of a 300–3400-cycle-per-second band-pass filter in the line circuit, and the use of a specified amount and kind of room noise at the receiving end. In addition, the particular adjustment of the circuit loss for which the articulation is X has to be specified when X has been determined.

It is not intended that this articulation reference system should be widely duplicated, but rather that the one system should be a means of co-ordinating in one laboratory the national articulation reference systems of all administrations.

Each administration will choose ordinary telephone instruments of its own type and will decide on typical connection circuits and typical (or limiting) local lines. Such an assembly of equipment constitutes a transmitting or receiving system to which transmitting and receiving articulation reference equivalents can be assigned by comparison with the articulation reference system in Paris.

For this purpose, a transmitting system to be tested replaces the transmitting portion of the articulation reference system. A receiving system similarly replaces the receiving portion of the reference system. One of the first requirements that will have to be fulfilled is that the sum of the articulation equivalents for the transmitting and receiving systems determined separately shall be equal to the articulation equivalent determined when both are simultaneously substituted for the reference terminal systems.

When a national articulation reference system has been thus calibrated, the administration to which it belongs will be able to set it up with the appropriate amount of attenuation to reproduce the reference condition X, and other arrangements of local line and connection circuit or other sets can be rated by comparison with the national articulation reference system in the same manner as this itself was rated in Paris.

Proceeding on these lines, substituting any equipment to be rated for the corresponding part in the national articulation reference system and readjusting the attenuation until the articulation is the same over the test system as over the unchanged reference system, it will be possible to build up a complete system of articulation reference equivalents for different combinations of sets, local lines, connection circuits, and inter-office trunks.

Further steps are necessary before these equivalents can be described as transmission-service ratings, which is the form in which they are to be used. It is necessary to determine by associated laboratory and service observations, (*a*) the loudness of sidetone for various combinations of local line and inter-office trunk, (*b*) the dependence of the transmitter output in service upon the loudness of sidetone, and (*c*) the average room noise in service and its influence on transmitter output and received quality if it is different from the room noise used with the reference articulation system.

With this information, corrections can be made to the articulation equivalents to allow for differences between transmitter outputs in service and in the articulation tests, and for differences, if any, between the average effects of room noise at the listening end in service and in the articulation tests. The articulation equivalents thus corrected become transmission-service ratings and can be used for planning purposes provided they approximately fulfil the requirement that addition of the ratings for all the parts of a circuit agrees with the rating determined for each circuit as a whole.

In carrying out such a programme as this there is, of course, much scope for interpolation of results by calculation of relative values of articulation, attenuation, and sidetone; it is to be expected that several subsidiary studies will be made in order to perfect or verify such methods. It may, for example, prove possible to calculate the differences in volume loss, distortion loss, and sidetone attenuation between the national articulation reference circuit and other commercial circuits using the same apparatus, so long as the differences are not too great. It will be necessary, however, in every case to establish the relation between sidetone intensity and transmitter output under service conditions.

8.1 LINE NOISE

So far, no consideration has been given to line noise. As line noise interferes with received speech by masking the weaker components, it is akin to room noise. The total noise interfering with speech is the resultant of the line noise and the room noise received by leakage and by sidetone; if the line noise and room noise components differ by 5 decibels or more the resultant is practically equal to the greater. It will be necessary, therefore, to introduce corrections for line noise when it is of the same order as, or greater than, the room noise entering the telephone ear. From the nature of the case, it is apparent that the total noise must be ascertained for the purpose of assigning noise impairment and that line-noise impairment cannot be separately assigned as an independent impairment.

Two courses are possible: either a fixed value of line noise must be used in the articulation reference system as part of the reference conditions, or a separate study of line-noise impairments must be made by each administration as part of the programme of establishing transmission-service ratings.

The first alternative has not been adopted, primarily because it was desired to make the Comité Consultatif International Téléphonique laboratory tests as simple as possible. The second alternative has the advantage that each administration will be able to discover the importance of line noise in its own system and direct its efforts towards the reduction either of sidetone or of line noise according to the indications of the results obtained.

In order that all interfering noise may be accounted for, line noise must be understood to include induced noise of all kinds, switching noise, and electrically transmitted noise from the far end.

9. *Application of Transmission-Service Rating*

The impairments that must be taken into consideration in determining the transmission-service rating of a circuit are of three principal types:

A. Those directly affecting received level, namely the overall attenuation (or reference equivalent) and the average transmitter output determined by the subscriber's reaction to sidetone and room noise.

B. Those affecting the useful received level by masking, namely line noise and room noise (direct and via sidetone).

C. Distortion, namely non-linear and amplitude distortion and frequency distortion of apparatus and lines including low-pass filters.

The total effect of these impairments when international and national standards of transmission were fixed on the reference-equivalent basis was considerably greater than it is to-day, and it is because continued planning on the old basis would fail to capitalize these improvements that the new conception of rating has developed.

For example, 25 years ago quite a large number of subscribers' sets were not designed to reduce sidetone as much as possible, the transmitter and receiver characteristics were less satisfactory than is now possible, and many long-distance lines had a cut-off about 2600 cycles per second or lower compared with 3000 cycles per second or higher to-day.

The reduction of impairments attributable to these improvements of the last quarter century is not a fixed amount applicable to all cases, but as an illustration the following figures are typical:

Improvement	*Reduction of Impairment in Decibels*
Anti-sidetone circuit:	
Louder speaking by subscriber[2, 5]	4.0
Reduced effect of room noise[2, 5]	2.0
Better transmitter and receiver frequency characteristics	5.0
Cut-off increased from 2600 to 3000 cycles per second	2.0
Total	13.0

None of these contributions to the transmission performance of the circuit would be taken into consideration if circuit rating continued on the basis of reference equivalents (volume rating). The subscriber would of course benefit from the improvements introduced, but sound economic planning demands that improvements should not automatically and accidentally go to the subscriber as transmission-service improvement but should be properly utilized to give the subscriber the benefit of both cost reduction and progressive improvement in transmission service.

Suppose, then, that economic study indicates that the 13-decibel reduction of impairment ought to be divided as 7.0 decibels improvement in transmission service and 6.0 decibels reduction of cost. The problem is how to design circuits for modern instruments with 6 decibels more attenuation than was allowed for the old instruments, 25 years ago.

The most economical layout of an exchange area network consistent with transmission requirements is obtained when the smallest practicable gauge of cable is used for all lines up to the length at which the permissible transmission loss for the area is reached; longer lines use the the next larger gauge of wire up to its transmission limit and still longer lines use a still larger gauge. In practice many other considerations interfere with this simple plan,[16] and local lines built up with two or more different cable gauges are common. In addition, there may be limitations imposed on the maximum line resistance by signalling requirements.

It is evident that in every exchange area there are a number of lines that have less than the maximum permitted loss. Consequently a revaluation of the permitted loss on the basis of transmission performance, allowing an increase of attenuation loss in the local lines, will permit cable plant economy to be effected only in respect of those lines which on the new assessment are transferred to the zone covered by a smaller gauge of cable.

In a particular case where the allowed transmitting loss for the area is 10 decibels on the volume basis, an increase in the permitted loss to 16 decibels (equivalent to 10 decibels on the new basis) would alter the maximum allowable length of each gauge of conductor as in Table 1.

Corresponding data[10] relating to the instruments used in this country show a similar tendency to increased resistance of subscriber's lines, and in such cases signalling limits may prevent the full realization of transmission gain as cost reduction.

Part or all of the additional loss may, however, be taken up in the inter-office lines, more particularly the loaded lines, because these, having practically flat frequency characterisitcs, have substantially the same rating on the performance basis as on the volume basis. On the other hand, non-loaded lines will on the performance basis (loss at 1600 cycles per second) be assigned worse ratings than on the former basis of loss at 800 cycles per second, and will therefore absorb some of the gain accruing from the lower losses assigned to local lines. In future, loaded lines can be run with lighter gauges of cable. If the local network and junction lines are in cost equilibrium, it is to be expected that the economies to be made by the application of transmission-service rating should be shared between them, but it is also to be expected that the cost-balance conditions will be changed because of the appreciation in performance rating of loaded lines.

The application of transmission-service rating has been studied in this country for some years; the results have been related to a particular system which will no doubt become what has been called in this paper a national articulation reference system.

The methods used and some of the conclusions reached have been recently published[6, 10] and need not be reviewed here. In summary, it appears that for fully exploiting economies in long-term planning, which should result from the introduction of performance rating, it is desirable to do three things:

A. Eliminate instruments with poor performance, which would have disappeared by now, but for the war.

B. Introduce signalling relays capable of working through lines of higher resistance.

C. Develop practical methods of jointing cables of smaller gauge than 6½ pounds.

TABLE 1

Gauge Lb/Mile	Volume Rating = 10 db		Transmission-Service Rating = 10 db	
	Max. Length Miles	Max. Resistance Ohms	Max. Length Miles	Max. Resistance Ohms
20	5.4	475	7.6	670
10	3.6	635	5.4	950
6½	2.6	707	4.0	1090
4	1.6	700	2.8	1230

In other European countries, steps have not yet been taken towards the practical introduction of performance rating; in new or reconstructed areas there will be good opportunities for realizing economy with the new method of rating at an earlier date than in old established areas, but in general the introduction of the new rating method cannot be expected to do more than prepare the way for economies which will mature in 10 to 15 years.

In considering the economic gain to be derived ultimately by the application of transmission-service rating, it is necessary to have a clear conception of the relation between the two main types of impairment, namely distortion impairments, and attenuation or equivalent impairments, which in sum may be called gross attenuation. When circuits are rated by transmission performance, reduction of distortion can be offset by increase of permitted attenuation, but whereas increase of attenuation always implies reduction in cost, reduction of distortion does not necessarily increase the cost. Hence, as distortion is reduced, the permitted (volume) transmission loss tends to be increased; as there is a practical limit to volume loss, distortion impairments (i.e., the decibel equivalents of given distortions) become smaller as the volume-loss limit is approached. The economic gain to be secured by the introduction of telephone instruments with less distortion is therefore smaller in systems designed for a relatively high toll terminal loss (volume basis) than in those planned for a lower loss.

10. Other Methods of Assessing Circuits

Space does not permit discussion of the various laboratory methods that are used for estimating the transmission quality of communication circuits; most of these have been studied with a view to their possible use to measure transmission performance.

Apart from various forms of articulation test, special interest attaches to "immediate-appreciation tests" introduced by W. H. Grinsted[11, 12] in 1937. In this test, listeners are merely asked to report whether they have clearly and immediately grasped the meaning of sentences read to them; herein seems to be contained the essential element that determines whether repetitions will be required in service. The test has many advantages especially as regards simplicity, but there are concomitant disadvantages in greater difficulty of standardizing the procedure and of unifying results.

Amongst other contributions[14] to the subject of quality rating, mention can be made only of those not published in English. K. Braun[14] has deduced, on the assumption that repetition rate is intimately related to sentence articulation, that small repetition rates are linearly related to sound or sentence articulation. He then defines a quality loss as $\Delta A/A$, a fractional loss in sound articulation which he finds can be related to a real or equivalent attenuation value b (nepers) by a formula of the type $\Delta A/A = \exp K(b - p)$, where K is a constant, b is attenuation and p is a constant for the circuit, depending upon the location along the attenuation axis of the falling part of the articulation curve.

On these foundations it is shown that quality losses can be combined as exponential functions of the attenuations, a conclusion that is useful mainly for indicating that the resultant quality loss in a given case is very largely determined by the factor individually responsible for the greatest loss.

Braun concludes that variations of line attenuation are so far from being a dominating variation in transmission quality that it is questionable whether it is justifiable to place such close limits on line attenuation as is usual.

Applying Braun's conclusion in another way it may be concluded, as already remarked, that circuits having high attenuation cannot be substantially improved in performance by correction of distortion.

A rather different approach to the rating of circuits was described in 1942 by Strecker and Von Susani,[15] who used judgment tests to determine the most favourable conditions for communication circuits. Two teams, each consisting of five persons, were used, and each listener was asked to adjust the line attenuation of a circuit under test (*a*) to the best value, (*b*) to give the loudest bearable reception, and (*c*) to give the quietest reception that was adequate without causing effort to the listener. The results of the tests were presented as statistical distribution curves, from which, by a sweeping generalization, it was concluded that the best value for the reference equivalent of a telephone circuit was one neper and that two nepers should not be exceeded. (The maximum value for international calls is fixed at present at 4.6 nepers.) The authors then proceeded to study the effect of noise and sidetone on the judgments of quietest acceptable reception and found that, by suitably choosing the reference condition in each case, the impairments due to room noise, line noise, or sidetone could be approximately expressed by a common curve corresponding to Table 2, which has been compiled from the curves in the original paper.

TABLE 2

Room Noise,* Sidetone Level† or Line Noise‡ in Phons or Decibels	Impairment in Decibels
−20	−0.9
−10	−0.9
0	0
15	8.7
23	17.4
30	26.1

* Room noise in phons relative to 40 phons.

† Sidetone level in decibels relative to a reference equivalent of 2 nepers (17.4 decibels).

‡ Line noise in decibels relative to 1.5 millivolts.

Such a generalization as this is not unreasonable since the impairments concerned are all due to noise interference.

An interesting conclusion reached by Strecker and Von Susani is that improvement of the frequency characteristic of a telephone set by eliminating resonant peaks does not, other things being equal, appreciably change the maximum permitted attenuation; it does, however, reduce the tolerated minimum attenuation.

One other suggestion that has been put forward must be mentioned: in 1938, the German administration presented results of simultaneous observations of repetition rate and average duration of call for varying amounts of circuit attenuation; they found that as the attenuation increased the repetition rate rose and the average duration of call decreased.

The author of the present paper made similar observations on factory calls in service and found a very regular reduction of call duration with decreasing sidetone. The dependence of call duration upon circuit conditions is, if it exists, important for traffic studies, and because it is easy to observe it might be of some assistance for transmission-service rating. The reality of the effect has yet to be established, the author's own results requiring further confirmation for the following reason, which is recorded to illustrate the type of pitfall in the path of those who make service observations.

The calls observed were trapped at random and put through the experimental equipment where the sidetone was adjusted; the subscriber therefore had no opportunity of becoming familiar with the abnormal condition presented, but on the contrary noticed when his call had been trapped that his set was working abnormally. It is thought that the observed reduction of call duration might have been due to a high proportion of calls in which the subscriber considered he had "a bad line" and either broke down the call to dial again or shortened the call because he was too much aware of its abnormality. This suspicion is based upon the fact that observations at the lowest possible level of sidetone had had to be discontinued because of subscribers' complaints.

11. *Conclusion*

The facts and circumstances relating to the solution of a problem, which is at present in a fluid state, have been presented. In the United States, effective ratings have been in use for 15 years or more; the keystone of the system there used is the dictum that circuits which give rise to equal repetition rate with the same group of subscribers are equal in effective rating.

In Europe, the keystone has been cut to the American pattern but the arch has not been built; the principle has been accepted without the practice. The reason for this state of affairs is the variety of conditions to be found in the different countries, making it necessary for each country to carry out a very extensive and difficult programme of tests; in each country independent data would have to be collected with very little mutual assistance from the data of another country, and finally the co-ordination of all the data into a homogeneous international system would be a formidable task, especially considering the difficulty of centralizing comparative tests under a single authority such as the Comité Consultatif International Téléphonique laboratory.

The American Telephone and Telegraph Company are in the happy position of the designers of a new car who have been able to gauge the public requirements by extensive trial sales to their employees of cars presenting many different features.

Other car designers must proceed without so much assurance of success, but they are well aware that the consumer public will be their judge in the long run and that the sales index is the measure of their success in service. How then will the designer proceed? He will estimate without exact knowledge what the customer requires, and how the customer will use his machine; he will allow margins or factors of safety in the components that may receive rough usage; he will consider the sales value of different features in relation to the cost, and, in general, his design tools will be laboratory, engineering, and cost data.

The planner of telephone networks must proceed on analogous lines; his transmission-service ratings must be based on laboratory data, illuminated by the knowledge of those factors that repetition-rate observations have shown to be important, and reasonable margins or factors of safety must be allowed in recognition of the imperfection of the methods used. A procedure such as this cannot be regarded as the final development of planning methods, because it does not give full recognition to the place awarded by the world's technical telephone conscience to repetition rate as the best known measure of transmission service.

More research is needed; to repeat the fundamental research in each country presents difficulties and might well be a misdirection of research effort. Effort directed towards elucidating the connections between repetition rate, objective design factors, and subscriber behaviour may add more to basic knowledge of transmission service than can be acquired by repeating the initial research in different countries each with its appropriate objective circuit elements. The former course aims at analysis of a fundamental nature, the latter at the piling up of data which would be difficult to co-ordinate and which in view of their empirical nature would have to be redetermined from time to time.

Nevertheless, the observation of repetition rate is so far the last word in measuring transmission quality in service, but, like other primary standards, recourse to it will be less frequent as experience and knowledge perfect the appropriate laboratory techniques and increase confidence in the results they yield. It is, however, impossible to avoid the conclusion that repetition-rate observations will have to be made on a certain scale in several countries if transmission-service rating is to be established on firm foundations.

Finally, the author wishes to emphasize the view that, in harmony with the economic principles stated at the beginning of the paper, it is the quality of the service given to all the subscribers that should be measured and controlled in the ultimate plan; designing the transmission network for the worst condition and evaluating performance in this worst condition should be regarded as a provisional step. It is not unknown for a telephone instrument with a peaky characteristic to give better performance through high attenuation than an instrument with a smoother characteristic which is superior on shorter circuits. In general, it is doubtful whether transmission-service rating will show advantages over volume rating where the system has been designed for the greatest possible volume loss; it is only when a good grade of service in terms of volume rating is given that the economic value of improved quality can become apparent. Extending the argument, transmission-service ratings based on limiting transmission conditions are likely to show smaller advantages than those which assess the average service given to all the subscribers.

The values of transmission-service ratings will always be, not only relative to, but dependent upon the reference conditions.

12. *Acknowledgments*

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14. *Appendices*

14.1 RELATION BETWEEN IMPAIRMENTS AND REPETITION RATE

Let x, y, $z \ldots .$ be circuit parameters which define the transmission properties of a circuit. Let R stand for repetition rate, which by hypothesis is a function of the circuit parameters.

$$R = F(x,y,z \ldots .). \tag{1}$$

Proceeding by variation of one variable at a time, it is possible to find experimentally

$$R_x = F(x,y_0,z_0 \ldots .) = F_x(x), \tag{2}$$
$$R_y = F(x_0,y,z_0 \ldots .) = F_y(y), \tag{3}$$
$$R_z = F(x_0,y_0,z \ldots .) = F_z(z), \tag{4}$$

where x, y, z are variable and measured from fixed reference values x_0, y_0, z_0.

If x stands for circuit attenuation, (2) represents a graph of repetition rate against attenuation, which may be used to evaluate as an equivalent attenuation the impairment due to variation of any other parameter from its reference value, e.g., the impairment for the value y is the value of x in (2) for which $R_x = R_y$, or if F_x^{-1} is the inverse function of F_x so that $F_x^{-1}(R) = x$, the impairments I_x, I_y, I_z due to the values x, y, z are in decibels:

$$\left.\begin{aligned} I_x &= x, \text{ by definition} \\ I_y &= F_x^{-1}(R_y) \\ I_z &= F_x^{-1}(R_z), \end{aligned}\right\} \tag{5}$$

which is a formal statement that I_y, for example, is the number of attenuation units by which x_0 must be changed to produce the same repetition rate as is observed when y_0 is changed to y.

In order that the impairments thus obtained, in decibels, may be useful, they must possess additive properties. If x_0, y_0, z_0 are changed to x, y, z in one circuit so that

$$R_{xyz} = F(x,y,z), \tag{6}$$

then it is necessary that the relation

$$F_{xyz} = F_x(x + I_y + I_z) \tag{7}$$

shall be satisfied.

When this relation holds $(x + I_y + I_z)$ is the impairment due to x, y, z relative to the reference conditions.

14.2 DISTORTION PARAMETER

In Section 3.1, it has been noted that the circuit parameters used to describe the physical characteristics of a circuit should be so defined that they are independent of each other. No difficulty occurs in defining and measuring the first four parameters listed in Section 3.2, but distortion is in a different class. It is necessary to include in this one parameter the effects of amplitude and non-linear distortion, for which no appropriate objective measurements have yet been found, as well as frequency distortion, which presents difficulties in objective measurement, and which when so measured is not expressed as a significant single figure.

The natural course is to seek a statement of distortion in terms of articulation, but articulation is considerably affected by received level and noise so that it requires a special definition to secure independence of the other parameters.

Such a definition can be made (at least in principle) if the distortion of a circuit is defined as the articulation measured when all the other parameters of the circuit have fixed reference values and the speaking level is standardized at the level appropriate to such a circuit. As a consequence of this definition, circuits which give the same articulation under these special conditions have the same distortion factors.[2]

Let D be the value of the distortion parameter for a circuit, and let D_0 be the reference value; let x and x_0 stand for variable and reference values of attenuation with the notation of Section 14.1, omitting explicit reference to parameters other than x and D, which are to be held at their reference values

$$R_d = F(x_0, D). \tag{8}$$

The impairment equivalent to D is

$$I_d = F_x^{-1}(R_d) \tag{9}$$

or $\quad I_d = x$, when $F(x, D_0) = F(x_0, D)$.

This provides the data for fixing the impairments to be ascribed to different values of distortion studied by repetition-rate counts.

14.3 CALIBRATION OF A NATIONAL ARTICULATION REFERENCE SYSTEM UNDER THE INTERIM PLAN

When the master articulation reference system has been fully defined, it will consist of a transmitting system, a fixed distortionless line with specified attenuation (with filter) and a receiving system. At the transmitting end, voice level and talking distance will be specified; at the receiving end, room noise will be specified.

In order to retain appropriate numerical continuity with reference equivalents (volume ratings) used in the past, numerical values of articulation reference equivalent (not necessarily zero) will be attributed to the transmitting and receiving ends of the articulation reference system, and an overall articulation equivalent in decibels will be assigned to the whole system.*

Thus there will be values of reference fixed as follows:

T_0 = convenient figure attributed to the transmitting system of the European articulation reference system and chosen so that transmitting-service ratings of commercial systems are comparable in magnitude with their reference equivalents.

R_0 = convenient figure attributed to the receiving system of the European articulation reference system chosen on the same grounds as T_0 *mutatis mutandis*.

N_0 = convenient figure for the articulation equivalent of the European articulation reference system such that transmission-service ratings of existing systems are comparable in magnitude with reference equivalents.

Careful choice of references as described seems absolutely essential especially during the period (necessarily rather long) in which the old system of ratings is in use simultaneously with the new system.

There will also be fixed a certain amount of attenuation for the line of the European articulation reference system; this will be denoted by N_k, and, since the basis of articulation equivalents is the "added loss" method, it will include an amount of attenuation L for this loss. Then,

---

* It is important to decide which particular transmission system shall have the same numerical value of transmission-service rating as its volume rating or reference equivalent.

$$T_0 + (N_k - L) + R_0 = N_0. \quad (10)$$

The set-up for the calibration of a national transmitting system is shown schematically in Fig. 5.

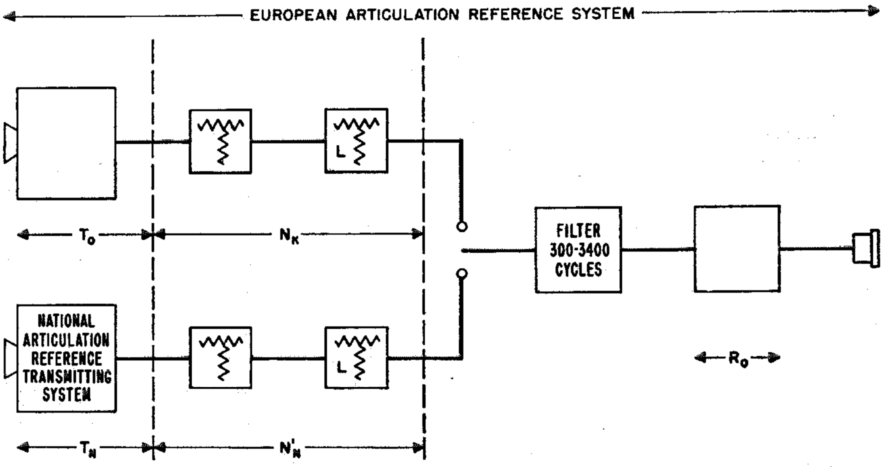

Fig. 5—Calibration of a national articulation reference transmitting system.

The value of $(N'_N - L)$ is to be determined by trial so that the articulation is the same over the two circuits. Then the national articulation reference transmitting system is worse than the European articulation reference transmitting system by $(N_k - N'_N)$ decibels, and is to be assigned the articulation equivalent

$$T_N = T_0 + N_k - N'_N. \quad (11)$$

The arrangements for calibrating a national receiving system and for making the overall check are shown in Figs. 6 and 7, it being understood that N''_N and N'''_N, respectively, are adjusted until the articulation is the same on the two systems compared.

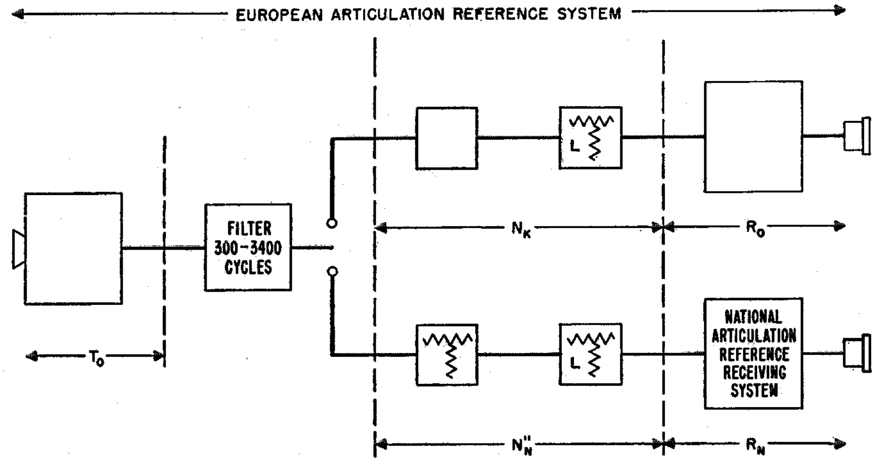

Fig. 6—Calibration of a national articulation reference receiving system.

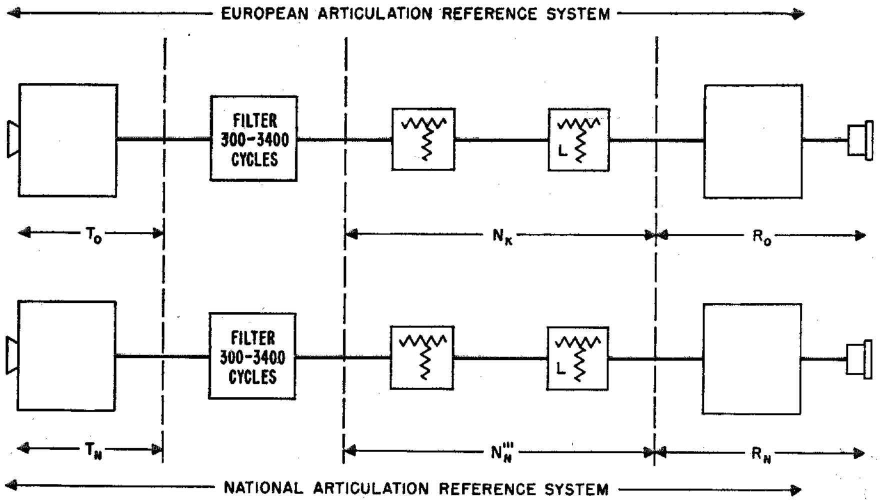
EUROPEAN ARTICULATION REFERENCE SYSTEM
FILTER
300-3400
CYCLES
L
T_0
N_K
R_0
FILTER
300-3400
CYCLES
L
T_N
N'''_N
R_N
NATIONAL ARTICULATION REFERENCE SYSTEM

Fig. 7—Adjustment of complete national articulation reference system.

The articulation equivalent of the national articulation reference receiving system is

$$R_N = R_0 + N_k - N''_N. \qquad (12)$$

The articulation equivalent of the national articulation reference system (combined transmitting and receiving system) is

$$T_N + R_N = N_0 - (N'''_N - L) \qquad (13)$$

or, by direct addition of (11) and (12),

$$T_N + R_N = T_0 + R_0 + 2N_k - N'_N - N''_N. \qquad (14)$$

From (13) and (14), substituting for N_0 from (10),

$$N_k = N'_N + N''_N - N'''_N, \qquad (15)$$

which condition must be satisfied if addition of the articulation reference equivalents of transmitting and receiving systems is to be valid.

The fixed amount of attenuation to be used as the line of the national articulation reference system when set up for use is, of course, N'''_N.

If (15) is not satisfied, better agreement must be sought by using a different reference adjustment of the articulation reference system, or a more complicated criterion of equality such as equal average articulation over a range of attenuations with or without weightings.

14.4 Determination of Transmission-Service Ratings under the Interim Plan

When a national articulation reference system has been calibrated in accordance with Section 14.3, it can be used to evaluate the articulation reference equivalents of other transmitting and receiving systems, in which different lengths or gauges of local lines are used. For this purpose the national system is set up and the system to be tested is compared with it in the same way as the national system was compared with the European articulation reference system. The appropriate values T_N and R_N are, of course, used instead of the articulation equivalents T_0 and R_0 of the European reference system.

For purely national purposes the amount of work to be done will be reduced by two-thirds if planning is done on the $T + R$ basis, i.e., the sum of the transmitting and receiving impairments is determined without evaluating them separately. In this case, the requirement of (15) will not arise, though it should nevertheless be met. For this procedure the only tests required are those analogous to Fig. 7. For international purposes, when the sets at the two ends are different, the necessity to determine T_N and R_N separately remains.

When the articulation equivalents have been determined for the local lines and gauges that are of interest, they have to be adjusted to transmission-service ratings by taking into consideration the differences between the articulation testing conditions and average service conditions.

The main considerations are sidetone and transmitter output. The national articulation reference system has necessarily some particular amount of sidetone which can be taken as the national reference sidetone. As the local line is varied, or the junction line-impedance is varied, the sidetone will change. It is necessary to establish by voice test or calculation, or by a mixture of both, how the attenuation of the sidetone path for speech changes from the reference sidetone value as the local line and trunk change. It is also necessary to determine, preferably by observations in service, the actual voice intensity at the transmitter for different amounts of sidetone attenuation, so that a curve can be drawn for voice level at the transmitter against sidetone attenuation relative to reference sidetone; the origin of the voice-level ordinates should be the value used in the national reference system under the conditions of articulation testing.

When this curve has been constructed, the transmitting articulation equivalent for any particular local line must be adjusted by the amount of the ordinate of the curve corresponding to the relative sidetone attenuation for that particular local line. When this has been done, the result becomes a transmitting-service rating.

Receiving articulation equivalents obtained by the procedure described will include the impairment due to reference room noise acting through the direct and sidetone paths, for the value of sidetone appropriate to a 600-ohm inter-office trunk. For other kinds of trunk and other values of room noise, the sidetone and the room-noise impairment will be different and appropriate corrections will be required.

When necessary, additional impairments due to line noise must be taken into consideration.

To complete the system of transmission-service ratings, it is necessary to have ratings for the different types of junction line.

Ratings can be found by substitution in the national articulation reference system, but the changes in reflection coefficients thus introduced make this a matter of some difficulty; however, studies already made indicate that the impairment per mile of junction can be evaluated as the attenuation per mile measured at a particular frequency in the neighbourhood of 1600 cycles per second. Reflection and interaction impairments can also be determined in theory by tests in the articulation reference system, but it is probably a sufficiently good approximation to treat these as purely volume losses, their values generally being of the order 1–3 decibels.

The effect of the junction impedance on sidetone, and hence on talking level and room-noise impairment, cannot be neglected, and must be taken into consideration as described above.

14.5 Field Data Relevant to Transmission-Service Rating

14.5.1 Variation of Transmitter Output for a Large Number of Subscribers

A series of observations was made in Bucharest on lines connected to public call boxes (see Table 3). Volume indicators were connected at the exchange end of the lines and additional lengths of cable were connected as required.

Another series of observations (see Table 4) was made on calls at Standard Telephones and Cables, Limited, at New Southgate, where the local lines were not altered but a range of variation of sidetone was secured by alteration of transmitter current and terminal impedance, with compensating amplification to avoid alteration of the transmission equivalent.

About 600 calls were observed for each condition.

Tables 3 and 4 both show that the spread of transmitter outputs or personal impairments at the transmitting end is not constant but tends to diminish, in the first case as the local line lengthens, in the second case as sidetone diminishes. As lengthening of the local line causes reduction of transmitter current and sidetone, the direct controlling cause may be the amount of sidetone

14.5.2 Speech Intensity at the Transmitter in Relation to Sidetone

The second series of observations, shown in Table 4, was made for the purpose of relating sidetone to speech intensity at the transmitter. Average readings of the volume indicator for

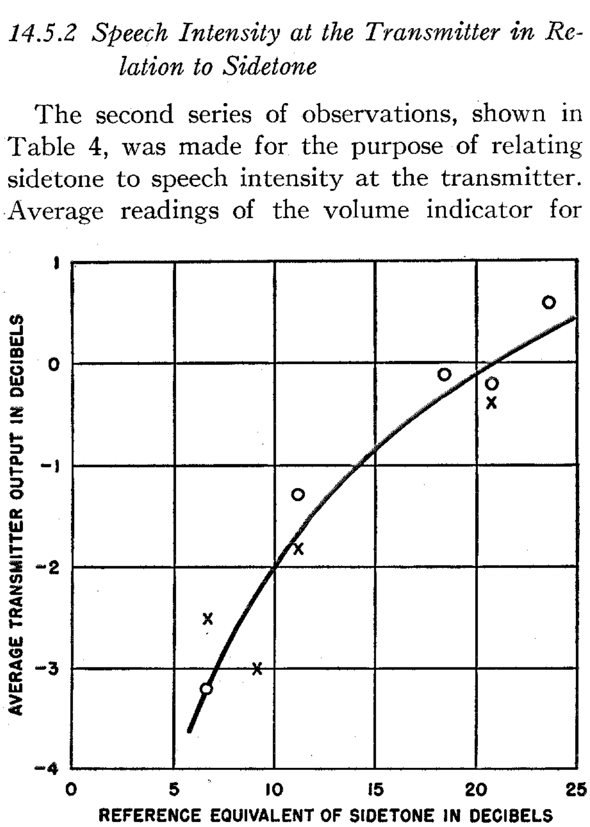

Fig. 8—Dependence of voice intensity upon sidetone.

TABLE 3

Added Local Line, 4-Pound Cable in Kilometers	Number of Observations	Standard Deviation of Volume-Indicator Readings in Decibels	Skewness
0	4000	7.96	−0.384
1	7200	7.56	−0.236
2	3200	7.45	−0.157
4	8000	6.77	−0.146
5	8000	6.47	−0.16

TABLE 4

Approximate Reference Equivalent of Sidetone in Decibels	Standard Deviation of Volume-Indicator Readings in Decibels	Skewness	Type of Connection
6.6	6.70	−0.28	Tie Line
6.6	5.56	−0.18	Internal
9.1	5.80	−0.05	Tie Line
11.2	5.16	+0.14	Tie Line
11.2	5.12	−0.29	Internal
18.5	4.56	+0.02	Internal
20.7	5.60	+0.06	Tie Line
20.7	4.66	−0.08	Internal
23.8	4.35	−0.13	Internal

each line condition were compared with the reading obtained for a standard speech intensity at a fixed distance. The results are plotted in Fig. 8.

It should be noted that the results in this Section are for the purpose of illustration, particularly of the range and variation in range of personal impairment. The telephone sets used were not the standard sets used in Great Britain.

14.6 ABBREVIATIONS, TERMS, AND DEFINITIONS

A. C.C.I.F.—the consulting international organization concerned with all aspects of international telephony.

B. A.T.&T.—the American Telephone and Telegraph Company, incorporating the Bell Telephone Laboratories.

C. A.E.N. in C.C.I.F. documents—see *H*.

D. Transmission performance—a general term for the degree of satisfaction with which a circuit fulfils its transmission function in service.

E. Effective transmission—transmission performance when considered in connection with repetition rate as the criterion.

F. Effective transmission rating—a numerical assessment (in decibels) of effective transmission, derived through the application of the criterion that circuits with equal repetition rate have equal effective transmission.

G. Transmission-service rating—a numerical assessment of transmission performance derived in any way understood or described.

(In earlier publications transmission service (rating) has been called effective transmission (rating); it has been decided by the Comité Consultatif International Téléphonique (in 1946) that the term "effective transmission" should be reserved for use as defined in *E* and *F*.)

H. Articulation reference equivalent (French A.E.N.—*affaiblissement équivalent de la point de vue de la netteté*) of a transmitting or receiving system or both together, is the difference between the attenuation of the line of the European or a national reference system, and the line of the system tested when both systems have the same specified articulation X.

(A distinction has been arbitrarily made between "equivalent" and "rating," the former is used in connection with laboratory articulation tests and the latter for final values to be used in planning as representative of service results.)

I. European articulation reference system—The European reference system modified by the inclusion of a 300–3400-cycle-per-second filter, and with room noise at the receiving end and a fixed specified attenuation in the line; used to determine the articulation reference equivalents of national articulation reference systems.

J. National articulation reference systems—commercial transmitting and receiving systems completed by a line containing a 300–3400-cycle-per-second cut-off filter and distortionless attenuation of such an amount that the articulation of the system is the same as that of the European articulation reference system.

K. Articulation reference value (X in the text)—the particular articulation value observed on the European articulation reference system or on a national articulation reference system; see *H*.

L. Transmitter output—used in the paper to express relative values of the subscriber's speech intensity at the transmitter, depending on his speaking intensity and the distance between his mouth and the transmitter; where necessary it is to be understood that corrections have been applied to eliminate the influence on the actual transmitter output of feeding current and line loss.

(Under service conditions "transmitter output" can be observed; the subscriber's voice intensity and speaking distance cannot be observed.)

M. Repetition rate—in principle, the average number of repetitions of words or phrases asked for in 100 seconds of actual conversation time; it is calculated from the total number of repetitions and the total time observed over a large number of calls. A more precise definition includes the detailed instruction given to observers as to what is a repetition and how the conversations are timed. Adverse comments are similarly counted per 100 seconds of conversation.

Position-Finding by Radio: First Thoughts on the Classification of Systems*

By C. E. STRONG

Standard Telephones and Cables, Limited, London, England

I WISH, first, to thank you very much indeed for the honour you have done me in electing me as your Chairman for this coming session. It is a great privilege to be the Chairman of the Radio Section, and certainly I shall be proud and happy to serve the Section to the best of my ability.

During the last two sessions, under the able guidance of the retiring Chairman, Professor Willis Jackson, and his predecessor, Mr. Mumford, we completed the task of committing to record the great store of new knowledge and technique which was an outcome of the war effort. That record is now in the pages of the *Journal;* it is not only a record of great scientific and engineering achievements, but also, we may well say, a monument to the skill and perseverance of many workers in the radio field.

In particular, we see how outstanding were the contributions made by the physicists, the mathematicians, and the chemists, who, coming in newly from other fields of work, applied themselves to our subject during the war. They brought in a fresh outlook and the latest scientific knowledge, and certainly also they made their mark in judgment and leadership. Being myself an engineer, it is not inappropriate for me, I think, to speak of the vitalizing effect of the new blood they introduced and to pay a high tribute to their valuable and important work.

The most spectacular achievements were in the fields of radar and radio aids to navigation, or, perhaps I should say more generally, in the whole art of position determination by radio. Here the widest new vistas were opened up and the most alluring prospects for the future were revealed. It remains to us now to apply this new knowledge to the needs of the day, and certainly it is a task to inspire interest and to fire enthusiasm.

But first, I believe that, to facilitate the orderly application of this knowledge, there is a need for us to rearrange our material and perhaps to overhaul our terminology. The art has moved so fast that we are almost overwhelmed by the flood of systems, and there has been little chance to sort them out in an orderly fashion and to see them in their proper relationships one to another. Some of the newer methods appear more fundamental than those with which we have previously been familiar, and that would seem to require some readjustment of our past ideas.

Now, however, with the completion of the record of war-time activity, to which I have just referred, we have the opportunity to stand back and survey the entire scene so that things may be viewed in their true perspective.

If, as a result, it appears that, in order to blend the new systems with the old, we shall require a new pattern of classification and some new terms, that will, of course, be a matter for consideration and agreement. It could not be carried far in an address such as this, yet it has seemed to me that I might venture to go a little way in that direction on a sort of preliminary reconnaissance.

I propose, therefore, to glance over the field of position determination by radio to obtain a first impression as to how systems are related to each other and, having in mind the table of classification given by Whelpton and Redgment in their paper at the Radiocommunication Convention, to consider some of the factors by which systems are distinguished.

First I will ask, What is the status of this art of position determination, or whatever it should be called, in relation to radio generally? Is it a part of radio or is it something different? Is it proper to use an expression such as "radio and radar devices," or would that suggest confusion of thought between the particular and the general? It is certainly a matter of the transmission of intelligence by radiation, and surely therefore it is a branch of radio. It involves techniques very similar to those employed in radiocommunication but it differs from communication in that the intelligence conveyed is restricted to geometrical information in contrast to the unlimited scope of intelligence conveyed in communication. It appears, indeed, to be on a par with communication as a main branch of the tree of which radio is the root and trunk. If that is so, what should it be called? The name must cover the whole field of direction and range determination. It must embrace direction-finding and radar, and much else. We speak of "radio aids to navigation," but that is cumbersome and not sufficiently general. I believe the title "radiolocation" would be ideal, though it has been used before in a different sense and it is perhaps somewhat awkward to borrow a term with such important historical associations. But it would be hard to find a name which could more aptly be applied as the comprehensive term to cover the whole of the field. The main branches of radio would then be radiocommunication and radiolocation, and we could speak of systems as communication systems or location systems.

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* Chairman's address to the Radio Section, Institution of Electrical Engineers, London, October 15, 1947. Reprinted from *Journal of the Institution of Electrical Engineers*, v. 95, Part I, pp. 31–35; January, 1948.

Accepting that for the time being, how should we proceed to divide up radiolocation itself? In view of the rapid development which has taken place, existing divisions are inadequate or ill-defined. Thus, while we have the term "direction-finding" to define parts of the field concerned with reception, we have no equivalent term to denote the process of transmitting azimuth information by means of appropriate beacon systems; and, while we make common use of the title "radar," its present meaning is far from clear. There is a lack, I think, of general terms to define main sections of the field, and in consequence we are often obliged to describe the general in terms of the particular by means of the code names of specific systems. There appears to be a need to examine the subject anew, starting from the firm ground of the common basic principles.

It might be well to look first to see if there is any lead to be obtained from what has occurred in the communication field. Here the first division is on the basis of purpose of application. We have telegraphy, telephony, broadcasting, television, and so on. Thereafter there can be further classification by the principal technical methods employed, such as modulation, whether variable-amplitude or constant-amplitude, or channel separation, whether by frequency separation or by time-sharing. But the various technical methods are common to the several functional categories, and so the division by function had to come first. The same will apply, I think, in radiolocation, and before drawing distinctions by technique we should first achieve some segregation of systems on a broader basis. It is more important, for instance, to see Gee and Decca first as closely related by function before dismissing them into widely separated camps because one is pulse technique and the other continuous-wave technique.

To make a start, we must go back to first principles and work from the general towards the more particular. The A.B.C. of the whole art is that by measuring the difference of the distances of an object from two fixed points we can find that the object is situated on a particular hyperbola, that being a locus of points of constant difference of distance from the fixed points: and by measuring the sum of the distances from the fixed points we can state that the object lies on a particular ellipse, being a locus of constant sum of the distances. In the first way I would say, taking some liberties with words, we obtain an idea of sideways displacement or direction and in the second an idea of outwards displacement or range. A single system of two spaced elements, with provision for measuring both the sum and the difference of the distances, could yield a fix by intersection of the hyperbolic and elliptic loci.

The distance relationships have, of course, to be found in terms of time of wave propagation. Let A and B (Fig. 1) be our two fixed points and C the mobile object. Then first, if we suppose there is a master transmitter at A, a repeater or slave transmitter at B and a receiver at C, by observing at C the difference in the arrival times of signals transmitted over the paths AC and ABC, and knowing the delay in AB, we can obtain the difference of the distances AC and BC and hence derive knowledge of the hyperbolic position line of C. This, of course, is the basis of such direction-giving systems as Gee and Decca. We could also reverse the process and transmit from the mobile C and have repeat and master receivers at B and A, respectively, so obtaining a hyperbolic direction-finding system as distinct from a direction-giving system.

Secondly, suppose we had a transmitter at A, a repeater at B, another repeater at C and a receiver at A; we could then measure at A the total propagation time round the loop AB, BC, CA, and knowing the delay AB, derive the sum of the transit times of the paths BC and CA, so finding the elliptic position line, or locus of constant time sums, on which C lies.

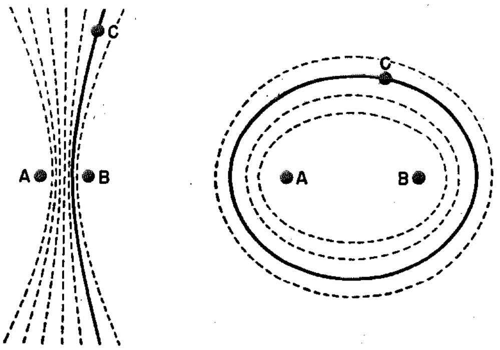

Fig. 1—Basic principles of radiolocation. (*Left*) Measurement of the differences of the distances AC and BC determines the hyperbolic position line of C. This is the general case of "direction" determination. (*Right*) Measurement of the sum of the distances AC and BC determines the elliptic position line of C. This is the general case of "range" determination.

These, as I see it, are the fundamental principles of radiolocation on which the whole art is based. They are not peculiar to hyperbolic and elliptic systems, as, in the special cases arising when the separation between the fixed points is small compared with the distance of the mobile object, they apply also to the classical direction-finding systems and beacon systems and to the usual methods of range measurements. Thus in elliptic systems, as the spacing is reduced the ellipses merge into circles, and when A and B coincide (Fig. 2) we have the normal conditions of range measurement when the distance measured is $AC + CA$, which is twice the range; and in hyperbolic systems, as the spacing is reduced the loci of constant difference of distance merge into straight radial lines.

The point I am making, therefore, is that all radiolocation systems are either hyperbolic or elliptic position-line systems, these including radial and circular position-line systems as special cases. The hyperbolic and radial systems are concerned with direction in a manner of speaking, and the elliptic and circular systems with "range."

From this it appears that the first division of radiolocation might be into two parts (Fig. 3), including on the one hand all time-sum methods giving "range," and on the other hand all time-difference methods giving "direction." If so, we should require eventually to coin new terms to denote these two main divisions of application, "X-dar" and "Y-dar," so to speak. For the present I shall refer to them by the terms "direction determination" and "range determination." These two categories would together cover the whole of the field, and subsequent steps would be a matter of further division within these categories.

Pursuing that, we must evidently provide, in the category of "direction determination," for transmitting and receiving systems. In general, the main methods can be applied in either way. Thus the Adcock direction-finding system has a transmitting counterpart in the omni-range beacon system; and, as has been mentioned, the principle of Gee could be applied to reception as well as to transmission. For the present, I shall use the terms "direction-giving" and "direction-finding" to denote these two functional divisions. The former would include Gee, Decca, omni-range beacons, overlapping beam course beacons, and all systems transmitting azimuth information, while the latter would embrace all the receiving counterparts of the above.

Then in both of the main categories it would be convenient to distinguish between broad-based systems and narrow-based systems. There is a considerable contrast in application between these two. In direction determination, for in-

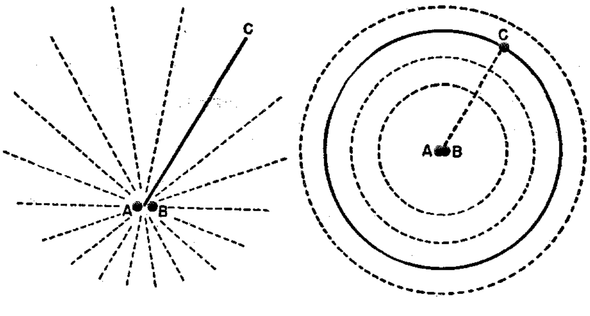

Fig. 2—Radial and circular systems are special cases of hyperbolic and elliptic systems. (*Left*) In the special case when the spacing AB is small compared with the distance AC, the difference of distances is $AB \cos \theta$ and the loci of constant difference of distance are straight lines. (*Right*) When the spacing AB is very small the loci of constant sum of the distances AC and CB are circles.

stance, the broad-based hyperbolic systems are applicable where it is important to carry as far out as possible the accuracy of position-line determination obtainable near in, while the narrow-based radial systems, on the other hand, are applicable when it is essential to obtain uniformity of indication all round, as in ordinary direction-finding and omni-range systems.

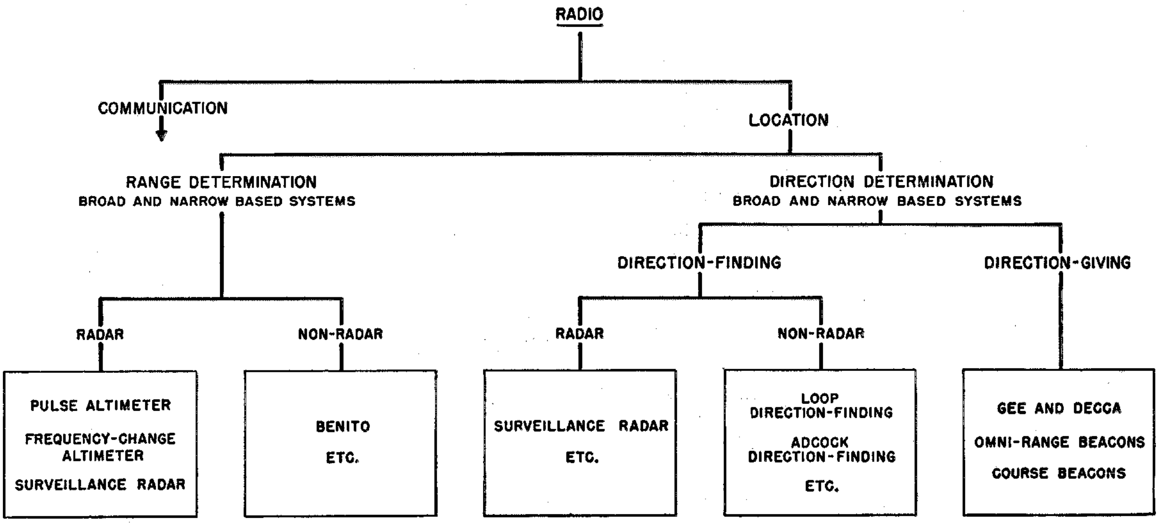

Fig. 3—Division of radio on a broad basis of function.

So far, we have not introduced radar into the family tree; but if, conforming, I think, to Sir Edward Appleton's definition, we mean by radar the radiolocation of objects without their active co-operation, then it is clear that in the categories of range determination and direction-finding there could be sub-division into radar and non-radar groups to differentiate between systems which do and those which do not require co-operation.

I have now, I hope, indicated a line of thought in the matter of a first division of systems by application or function. We are still left, however, with systems falling in the same categories which clearly require further differentiation. For example, while Gee and Decca would so far be classed together as broad-based direction-giving systems, there is a marked difference between them in technique. We have in fact reached the stage where further differentiation must be done according to the technical methods used. This is a large subject, and I shall not aim to do more than introduce it by some observations on three of the factors which I think would be relevant. These are, first, the method by which transmissions are marked so that differences in times of arrival can be observed; secondly, the method of modulation; and, thirdly, the method of channel separation.

Consider first the method of marking (Fig. 4); this is a question of the calibration of the yardsticks of radiation sent out. The method which, it seems to me, springs most naturally to mind is to start the transmission suddenly and observe the times of arrival of the sharp fronts. In practice, the transmission is marked by a succession of pips. This, of course, is the pulse method as used in Gee and primary radar. The transmission is in the form, we might say, of a "comb carrier," and time intervals are observed as the displacement between the teeth of the interlaced combs received over circuits of different delay.

Secondly, the transmission can be marked by the peaks and valleys of low-frequency sine-wave variation impressed on the radio-frequency wave. (I am trying to avoid using the word "modulation" in this connection, as no geometrical information is conveyed by this continuously repetitive variation.) Differences of distance are observed by phase comparison of the low-frequency envelopes, and the method can be referred to as the "tone-phase comparison method." It is exemplified in the Benito ranging system and in certain proposed course beacon systems.

Thirdly, we have the radio-frequency phase-comparison method in which the wavelength of the radiation itself is the unit of graduation, and differences of distance are found in terms of the number of whole wavelengths and the fraction of a wavelength by which one path is longer than another. This is exemplified in direction-finding omni-range beacons and in Decca.

There is a sharp distinction between the pulse method and the two phase-comparison methods, in that in the former the transmission is discontinuous and the method is therefore inherently appropriate to time-sharing methods of multiplex and to time discrimination between signals, while in the latter the transmission is continuous and separation of signals must be by difference of frequency.

I pass on now to remark on the second factor of technical method I mentioned, namely modulation. This term as applied to radiolocation should, I think, be taken as referring to the characterization of signals whereby geometrical information is conveyed. In that sense, it is the signal characteristic which changes with a change in the position of the mobile object. We have, as in communication, the two main kinds of modulation, namely variable-amplitude modulation and constant-amplitude time modulation, of which frequency modulation and phase modulation are varieties (Fig. 5).

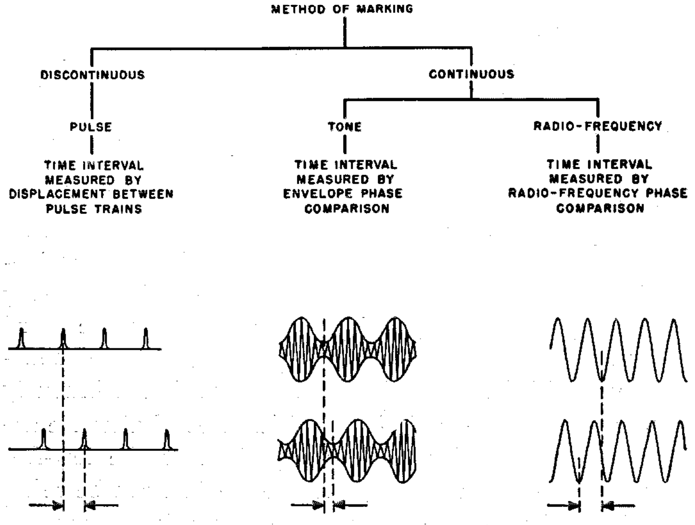
METHOD OF MARKING
DISCONTINUOUS
CONTINUOUS
PULSE
TONE
RADIO-FREQUENCY
TIME INTERVAL MEASURED BY DISPLACEMENT BETWEEN PULSE TRAINS
TIME INTERVAL MEASURED BY ENVELOPE PHASE COMPARISON
TIME INTERVAL MEASURED BY RADIO-FREQUENCY PHASE COMPARISON

Fig. 4—Division of radiolocation according to method of marking. Transmissions are marked to enable differences in times of arrival to be observed.

To take some examples, in overlapping beam systems the modulation is amplitude modulation, and the degree of modulation is the ratio of the amplitudes of the patterns. This changes with direction, and position lines are indicated accordingly. In Gee and in pulse-ranging systems the modulation is pulse-time modulation, i.e., the displacement in time of one train of pulses with respect to another train serving as a reference. It is a form of envelope phase modulation and is exactly the same kind of modulation as is used in some multi-channel pulse-communication systems. In the Benito ranging system the modulation is tone-envelope phase modulation, being the variation of the relative phases of two tone envelopes, and finally in Decca the modulation is radio-frequency phase modulation, being the change of phase of one radio-frequency signal with respect to another serving as a reference of phase.

While we have the same kinds of modulation as in communication, there is a very distinct difference in the way the modulation is impressed. Modulation of the signal with geometrical information in radiolocation can be effected only by virtue of the spacing of antennae. We might possibly use the expression "space modulation" in that connection.

To illustrate this point, consider a system of Gee character in which interlaced pulse trains of "comb carriers" with a certain fixed delay between them are fed to two radiators *A* and *B*. Imagine first that the radiators are brought together. Then a mobile receiver circulating round the system would receive the two carriers constantly with the initial fixed delay between them. The signal is unmodulated in the sense applying in radiolocation, and no position information is radiated. Now suppose we begin to move *B* away from *A*. We commence thereby to apply space modulation, and we have the situation that as the receiver rotates the *B* pulses received are cyclically advanced and retarded about their mean occurrence time with relation to the occurrence time of the *A* pulses. This is precisely what happens in a pulse-time-modulated communication sys-

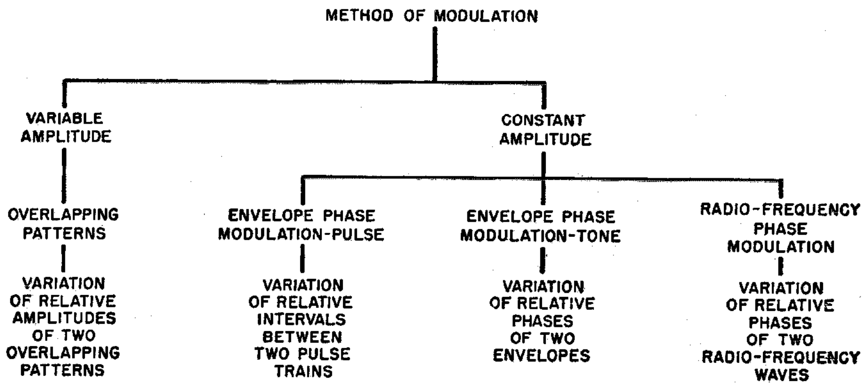
METHOD OF MODULATION
VARIABLE AMPLITUDE
CONSTANT AMPLITUDE
OVERLAPPING PATTERNS
ENVELOPE PHASE MODULATION-PULSE
ENVELOPE PHASE MODULATION-TONE
RADIO-FREQUENCY PHASE MODULATION
VARIATION OF RELATIVE AMPLITUDES OF TWO OVERLAPPING PATTERNS
VARIATION OF RELATIVE INTERVALS BETWEEN TWO PULSE TRAINS
VARIATION OF RELATIVE PHASES OF TWO ENVELOPES
VARIATION OF RELATIVE PHASES OF TWO RADIO-FREQUENCY WAVES

Fig. 5—Division of radiolocation according to method of modulation. In radiolocation, modulation is "space modulation," applied by the spacing of antennae.

tem when a channel is modulated, the channel pulses being advanced and retarded cyclically with relation to the train of synchronizing pulses. If the receiver is arrested in its circulation round the system a direct-current value of the modulation is observed which is indicative of the position line of the receiver.

The third point I mentioned is the method of channel separation. I think it can be said that radiolocation is fundamentally a multiplex proposition in that, in "range" systems it is required to transmit and receive simultaneously on the same site without interference, and in "direction" systems it is required to transmit simultaneously over multiple paths without interference, or, in the case of amplitude-modulated systems, to lay down separate overlapping patterns.

Two methods exist, namely "frequency multiplex" in which channels are separated in frequency, and "time-sharing multiplex" in which channels are separated in time (Fig. 6).

The distinction can be illustrated by comparisons between Gee and Decca and between pulse-ranging systems and tone-phase comparison-ranging systems.

In Gee, the transmissions from the spaced points are on the same frequency, but, being in the form of interlaced pulse trains, they are kept distinct by time-sharing. The signals pass through the same receiver, but because of the separation in time there is no interference. In Decca, on the other hand, the transmissions from the spaced points are on different frequencies and they are separated at the receiver by frequency-selective circuits (actually two receivers). The point is that in Gee the channels are kept distinct by time-sharing, exactly as they are in a multi-channel pulse-communication system, while in Decca they are kept distinct in the familiar way by frequency separation.

As between pulse- and tone-phase comparison ranging systems, the same distinction applies. In the pulse method, simultaneous (if we may call it so) transmission and reception on the same site is possible by virtue of the fact that the outgoing and incoming trains of pulses are interlaced in time. In the tone-phase comparison method, on the other hand, there has to be a frequency conversion at the responder at the remote point so that the incoming signals can be separated from the outgoing signals by being on a different frequency.

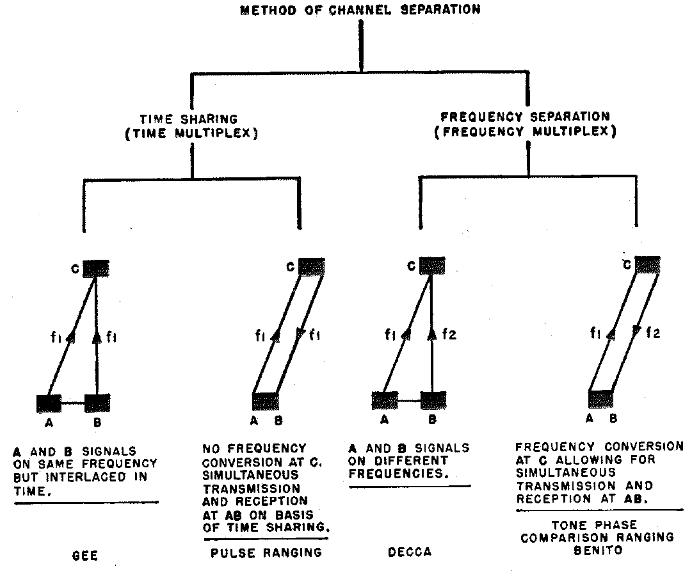
METHOD OF CHANNEL SEPARATION
TIME SHARING (TIME MULTIPLEX)
FREQUENCY SEPARATION (FREQUENCY MULTIPLEX)
C
f1
f1
A
B
A AND B SIGNALS ON SAME FREQUENCY BUT INTERLACED IN TIME.
GEE
C
f1
f1
A B
NO FREQUENCY CONVERSION AT C. SIMULTANEOUS TRANSMISSION AND RECEPTION AT AB ON BASIS OF TIME SHARING.
PULSE RANGING
C
f1
f2
A
B
A AND B SIGNALS ON DIFFERENT FREQUENCIES.
DECCA
C
f1
f2
A B
FREQUENCY CONVERSION AT C ALLOWING FOR SIMULTANEOUS TRANSMISSION AND RECEPTION AT AB.
TONE PHASE COMPARISON RANGING BENITO

Fig. 6—Division of radiolocation according to method of channel separation. In general, radiolocation involves multiplex operation.

The three factors of method I have briefly discussed, namely marking, modulation, and channel separation, are evidently closely related. Thus, where pulse-marking is used, time multiplex is obviously indicated, and indeed the main object of pulse-marking is to obtain time-multiplex effects, including the resolution of signals resulting from reflections from many objects, but when phase-comparison methods are used there is an incentive to use frequency multiplex to avoid the necessity of successive as distinct from continuous phase comparison. Yet the factors need some separate consideration, as there are cases in which the methods are associated in less usual ways, as for example in P.O.P.I., which, though a phase-comparison system, relies on time multiplex for channel separation.

The possible use of these or similar factors in the preparation of a complete chart of classification is something which must be left. I have aimed only at indicating some of the factors which would enter into the consideration. However, there is one point I would emphasize as having a bearing on our understanding of the term radar. That is the distinction, as applied to radar, between discontinuous and continuous marking, or, in more common terms, between pulse and continuous-wave techniques. Either method is applicable when we are concerned with the location of a single object, but when we are concerned with "seeing" very many objects simultaneously, as for example in marine or ground-surveillance radar, we are almost bound to resort to pulse technique, which enables us to work very many echoes simultaneously on what is, in effect, a basis of time-sharing multiplex operation. As both methods have their place, it seems that we should reserve the term radar for the art of radiolocation without co-operation, irrespective of the technical methods used, and avoid any tendency to associate the term particularly with pulse technique and time-sharing multiplex methods.

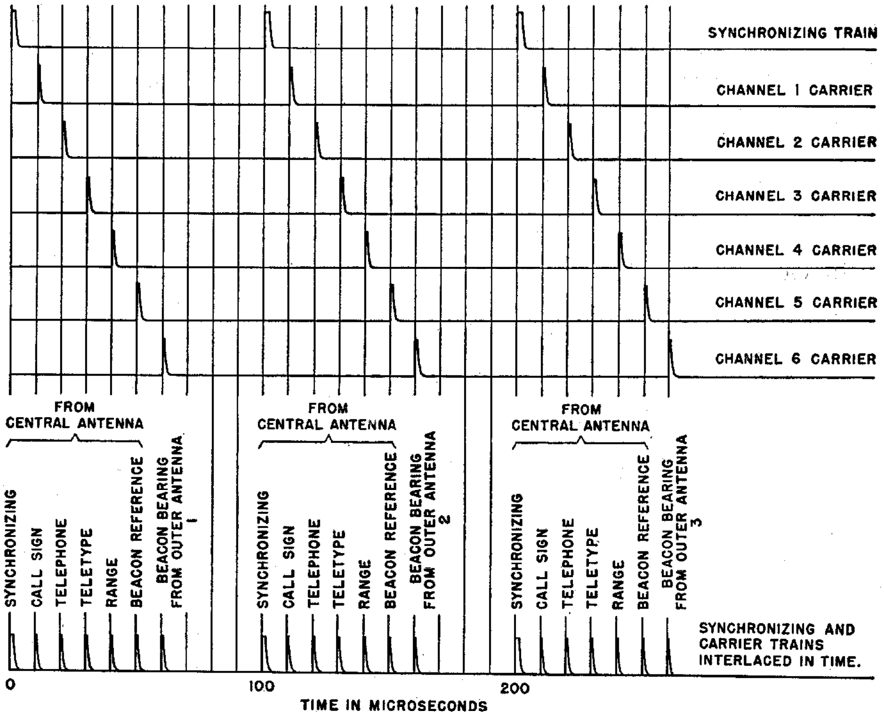

Fig. 7—Time-multiplex beacon and communication system. Distribution of carriers in time.

I have tended throughout in this brief review, I think, to emphasize the close relationship between radiolocation and radiocommunication as branches of a single art. Now in the time which remains, I propose to describe very briefly a system which, although somewhat hypothetical at present, will serve to illustrate that point. This is a time-sharing multiplex system combining radiolocation and radiocommunication services to aircraft on a common frequency.

The idea is that a multi-channel pulse communication transmitter, such as might be used for a point-to-point radio-link service, could possibly be used as an omni-range beacon while at the same time serving to transmit a number of communication channels. The transmitter delivers a number of interlaced pulse trains or comb carriers—one for synchronization and one for each channel (Fig. 7). The communication channels are modulated by advancing and retarding the pulses in accordance with the input speech or tone signals. The omni-range bearing channel is similarly pulse-time modulated, only in this case the modulation is space modulation, being effected by transmitting the pulses of the channel carrier successively from a number of antennae spaced evenly round the station on a circle of about 300-metre radius. The pulse train then becomes sinusoidally modulated owing to the different distances of the radiators from the receiver. The modulation is detected in the receiver, giving a low-frequency tone, the phase of which varies with the bearing of the receiver. By phase comparison between that signal and a reference tone transmitted on one of the communication channels the bearing can be determined.

In addition, the aircraft could measure its distance from the beacon. The aircraft transmits a suitable tone on its very-high-frequency equipment. The tone is received at the ground station and sent back on one of the communication channels, so that by phase comparison in the aircraft between the outgoing and incoming tones the range can be derived as in the Benito system. We should thus have a combined system giving azimuth, range, and communication, all on a single frequency on a basis of time-sharing multiplex operation.

And now to summarize very briefly: We are concerned with one art, which is radio, dividing, as I see it, into radiocommunication and radiolocation, applying the latter term now in a new and broader sense to embrace all methods of position determination by radio.

Then it appears that radiolocation itself might be divided initially into two sections concerned, respectively, with time-sum and time-difference measurement, giving in the one case elliptical and circular position lines and in the other, hyperbolic and linear position lines. For the time being, I used the terms "range determination" and "direction determination" to denote these two fundamental divisions.

Within these two categories there can be further division as between broad-based systems and narrow-based systems, and in "direction determination" we must distinguish between transmitting and receiving systems, or, as I have said, between "direction-giving" systems and "direction-finding" systems.

Further, in range and direction-finding systems we must distinguish between radar and non-radar applications, reserving the term "radar" for position determination without co-operation.

Then, having first classified systems on broad lines related to function, we could differentiate further by the technical methods used. Some of the factors which would be relevant in that connection are, first, the method of marking, whether discontinuous as in pulse systems or continuous as in tone or radio-frequency phase-comparison methods; secondly, the method of modulation, whether amplitude modulation as in overlapping-pattern systems or constant-amplitude time modulation as in systems depending on pulse-interval measurement or phase comparison; and thirdly, the method of channel separation, whether by time multiplex or frequency multiplex.

Finally, it will be understood that my approach to this subject has been purely exploratory. I have aimed only at framing a question and whetting the appetite for a wider discussion of the problems involved.

Simplified Procedure for Computing Behavior of Multiconductor Lossless Transmission Lines

By SIDNEY FRANKEL

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PROBLEMS involving radio-frequency transmission lines of more than two conductors, or problems involving combinations of two-conductor lines lying geometrically parallel, occur frequently. It is often useful to treat such lines as though they were lossless and operating in the *TEM* mode. Under such conditions, the description of the line behavior is particularly simple. It can be shown that, apart from terminal conditions, the electromagnetic conditions on the line can be expressed in terms of the velocity of propagation, which depends only on the electromagnetic constants of the medium, and on an additional set of constants, which may be either the coefficients of electrostatic capacitance of the line or the coefficients of external magnetostatic inductance.

* * *

1. *Line Equation*

Consider the case of a uniform lossless line consisting of $n+1$ cylindrical conductors of arbitrary cross section excited in the *TEM* mode. As a result of this excitation, two waves of fields travel along the line in opposite directions, accompanied by currents in the various conductors and potential differences between them. In the present discussion, the potential of the $(n+1)$th conductor is taken as zero and the potentials of all other conductors are referred to it. If we take rectangular coordinates x,y,z with the direction of increasing z as the direction of travel of the forward wave, then the forward-wave electric and magnetic fields are given in meter-kilogram-second units by

$$\left.\begin{aligned} \mathbf{E}_f &= \mathbf{E}_1(x,y)f(z-ct), \\ \mathbf{H}_f &= \left(\frac{\mathcal{E}}{\mu}\right)^{\frac{1}{2}}(\mathbf{K}\times\mathbf{E}_f), \end{aligned}\right\} \tag{1}$$

where

$\mathbf{E}_f$ = forward-wave electric vector,
$\mathbf{E}_1$ = a vector function of position,
f = a scalar function,
$\mathbf{H}_f$ = forward-wave magnetic vector,
$\mathcal{E}$ = dielectric constant,
μ = permeability constant,
$\mathbf{K}$ = unit vector in z direction,
$c = (\mu\mathcal{E})^{-\frac{1}{2}}$ = velocity of wave propagation,
t = time.

$\mathbf{E}_1$ depends on the conductor charge distributions in any transverse plane, while f depends on the way the line is excited, or may be considered as being excited, at any transverse plane as a function of time.

Similarly the back-wave fields are given by

$$\left.\begin{aligned} \mathbf{E}_b &= \mathbf{E}_2(x,y)g(z+ct), \\ \mathbf{H}_b &= -\left(\frac{\mathcal{E}}{\mu}\right)^{\frac{1}{2}}(\mathbf{K}\times\mathbf{E}_b), \end{aligned}\right\} \tag{2}$$

where

$\mathbf{E}_b$ = back-wave electric vector,
$\mathbf{E}_2$ = a vector function of position,
g = a scalar function,
$\mathbf{H}_b$ = back-wave magnetic vector.

Similar comments apply to $\mathbf{E}_2$ and g as to $\mathbf{E}_1$ and f, respectively.

By virtue of (1) and (2), simple relations exist between currents and charges on the various conductors. As the conductors are assumed to be perfect, $\mathbf{E}_f$ and $\mathbf{E}_b$ are normal to the conductor surfaces, while $\mathbf{H}_f$ and $\mathbf{H}_b$ are tangential thereto. Let $\mathbf{E}$ and $\mathbf{H}$ represent either forward- or back-wave intensities, and let q_k be the total charge per meter on the kth conductor. On a section of surface bounded by transverse planes one meter apart, Gauss' theorem yields

$$q_k = \mathcal{E}\oint \mathbf{E}\cdot d\mathbf{s}, \tag{3}$$

the loop integral being taken on the path $\mathbf{s}$ represented by the intersection of the kth conductor by a transverse plane.

Let I_k be the total current in the kth conductor. Then by Ampere's theorem

$$I_k = \oint \mathbf{H}\cdot d\mathbf{s}, \tag{4}$$

where the path of integration is the same as for (3). Equations (1) through (4) then lead immediately to

$$I_k = \pm c q_k, \tag{5}$$

the positive or negative sign depending on whether the wave under consideration is a forward wave or a back wave.

In a transverse plane, the electric field is described by a potential that satisfies Laplace's equation. Hence the charges on the various conductors are characterized by Maxwell's coefficients of capacitance and the conductor potentials by

$$q_k = \sum_{j=1}^{n} C_{kj} V_j, \quad k = 1, \ldots, n, \tag{6}$$

provided that

$$\sum_{j=1}^{n+1} q_k = 0,$$

where

$C_{kj} = C_{jk}, \quad j,k = 1, \ldots, n,$
$\quad$ = coefficient of capacitance between jth and kth conductor,
V_j = potential of jth conductor, $j = 1, \ldots, n$.

Thus, we have immediately

$$I_k = \pm \sum_{j=1}^{n} Y_{kj} V_j, \quad k = 1, \ldots, n, \tag{7}$$

where

$Y_{kj} = Y_{jk} = cC_{kj} = cC_{jk}, \quad j,k = 1, \ldots, n,$
$\quad$ = coefficient of characteristic admittance between jth and kth conductor.

For waves of a single angular frequency ω, the root-mean-square potential of the kth conductor may be written

$$V_k = A_k e^{-i\theta} + B_k e^{i\theta}, \; k = 1, \ldots, n, \tag{8}$$

where

A_k, B_k = arbitrary constants,
θ = electrical distance along the line,
$\quad = \omega z / c$.

The terms in the right-hand member of (8) correspond to the forward wave and back wave, respectively; A_k and B_k are usually expressed in terms of conditions at the ends of the line. Then (8) and (7) imply

$$I_k = \sum_{j=1}^{n} Y_{kj}(A_j e^{-i\theta} - B_j e^{i\theta}), \; k = 1, \ldots, n. \tag{9}$$

Equations (8) and (9) are the general steady-state solutions of the lossless multiconductor line. They reduce to the usual two-wire lossless-line equations when $n = 1$.

Equation (7) may be solved for the V_j's to yield

$$\begin{aligned} V_j &= \pm \sum_{k=1}^{n} Z_{jk} I_k, \\ &= \pm \sum_{k=1}^{n} c Z_{jk} q_k, \\ &= \pm \sum_{k=1}^{n} P_{jk} q_k, \end{aligned} \tag{10}$$

where the P_{jk}'s are recognized as Maxwell's coefficients of potential, and Z_{jk} may be termed a characteristic-impedance coefficient. Thus,

$$Z_{jk} = \frac{P_{jk}}{c}. \tag{10A}$$

The general line equations may also be written

$$\left.\begin{aligned} I_k &= C_k e^{-i\theta} + D_k e^{i\theta}, \\ V_k &= \sum_{j=1}^{n} Z_{jk}(C_k e^{-i\theta} - D_k e^{i\theta}). \end{aligned}\right\} \tag{11}$$

It is seen from these considerations that, apart from terminal conditions, the line behavior may be characterized completely by Maxwell's coefficients of capacitance and by the velocity of propagation. The well-known expression for the characteristic admittance of a two-wire line:

$$Y_0 = cC \tag{12}$$

is a special case of this result.

2. *Inductance Coefficients*

It may be shown that the line may equally well be characterized by its coefficients of inductance rather than by the capacitance coefficients; as a result, it is possible to compute the former from the latter.

The coefficient of external inductance L_{jk} is defined as the magnetic flux per unit length of line passing between the jth conductor and the reference $(n+1)$th conductor when the current in the kth conductor is one ampere. This leads to

$$\phi_k=\sum_{j=1}^{n} L_{kj}I_j \tag{13}$$

for the flux per unit length of line between the kth conductor and the $(n+1)$th conductor. Applying Faraday's law to a differential length of line then yields

$$\frac{\partial V_k}{\partial Z}=-\sum_{j=1}^{n} L_{jk}\frac{\partial I_k}{\partial t}. \tag{14}$$

For waves traveling with velocity c,

$$\frac{\partial}{\partial t}=\pm c\frac{\partial}{\partial Z},$$

the lower or upper signs depending on whether the wave is a forward wave or a back wave. Using this result in (14) and integrating with respect to Z yields for traveling waves

$$V_k=\pm\sum_{k=1}^{n} cL_{jk}I_k. \tag{15}$$

Comparing (15) with (10), we have

$$cL_{jk}=Z_{jk}=\frac{P_{jk}}{c} \tag{16}$$

or

$$L_{jk}=\frac{P_{jk}}{c^2}. \tag{16A}$$

On the other hand, as is well known,

$$P_{jk}=\frac{\Delta_{jk}}{\Delta},$$

where Δ is the determinant $|C_{jk}|$ and Δ_{jk} is the minor of the jth row, kth column of Δ. Thus, the L_{jk} may be computed when the C_{jk} are known.

It should be noted that since C_{jk} is proportional to $\mathcal{E}$, P_{jk} is proportional to $\mathcal{E}^{-1}$; therefore L_{jk} is proportional to $\mathcal{E}^{-1}c^{-2}=\mu$, so that the dielectric constant and velocity of propagation disappear in a calculation of L_{jk}.

An additional set of relations exists between the L's and C's,

Form the product $L_{jk}C_{kl}$ and sum over k, thus

$$\sum_{k=1}^{n} L_{jk}C_{kl}=\frac{1}{c^2}\sum_{k=1}^{n} P_{jk}C_{kl}=\frac{1}{c^2\Delta}\sum_{k=1}^{n} \Delta_{jk}C_{kl}.$$

By elementary theory of determinants, the last sum is zero for $j\neq l$ and is equal to Δ for $j=l$; thus

$$\sum_{k=1}^{n} L_{jk}C_{kl}=\frac{1}{c^2}\,\delta_j^l=\mu\mathcal{E}\,\delta_j^l, \tag{17}$$

where

$$\begin{aligned}\delta_j^l&=\text{Kronecker delta}\\&=0,\ k\neq l\\&=1,\ k=l.\end{aligned}$$

2.1 EXAMPLES

The usefulness of the results above will now be illustrated by some examples. These will be restricted to the case of two small round wires at equal heights above plane ground *in vacuo*, as shown in Fig. 1. For this three-conductor system, $n=2$. Approximate values for the admittance coefficients are well known. For D, $h\gg a$,

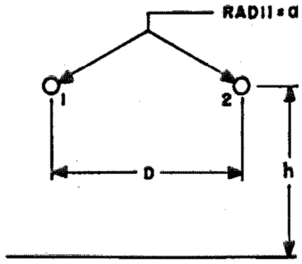

Fig. 1.

$$\left.\begin{aligned}Y_{11}=Y_{22}=cC_{11}&=-\frac{1}{60}\frac{\xi}{\xi^2-\eta^2},\\ Y_{12}=Y_{21}=cC_{12}&=\frac{1}{60}\frac{\eta}{\xi^2-\eta^2},\end{aligned}\right\} \tag{18}$$

where

$$\left.\begin{aligned}\xi&=\ln\frac{a}{2h},\\ \eta&=\ln\frac{D}{G},\\ G&=(D^2+4h^2)^{\frac{1}{2}}.\end{aligned}\right\} \tag{18A}$$

The usual characteristic admittances of the line may be computed from the admittance coefficients. To compute the characteristic admittance, we assume forward waves only so that (8) and (9) become

$$\begin{aligned}V_1 &= A_1 e^{-i\theta},\\ V_2 &= A_2 e^{-i\theta},\\ I_1 &= (Y_{11}A_1 + Y_{12}A_2)e^{-i\theta},\\ I_2 &= (Y_{21}A_1 + Y_{22}A_2)e^{-i\theta}.\end{aligned}$$

For a balanced line, $V_2 = -V_1$, $A_2 = -A_1$, $I_2 = -I_1$, and

$$Y_0 = \frac{I_1}{V_1 - V_2} = \frac{I_1}{2V_1} = \frac{Y_{11} - Y_{12}}{2}. \qquad (19)$$

For an unbalanced-fed line, $V_2 = V_1$, so that $A_2 = A_1$, $I_2 = I_1$, and

$$Y_0 = \frac{I_1 + I_2}{V_1} = 2(Y_{11} + Y_{12}). \qquad (20)$$

Using (18) and (18A) in (19) and (20) yields the well-known results

Balanced Line $\quad Y_{ob} = \left(120 \ln \frac{2hD}{aG}\right)^{-1}, \qquad (21A)$

Unbalanced Line $\quad Y_{ou} = \left(30 \ln \frac{2hG}{aD}\right)^{-1}. \qquad (21B)$

2.2 Input Admittance of a Short-Circuited Line Above a Ground Plane

The case to be analyzed is shown in Fig. 2.

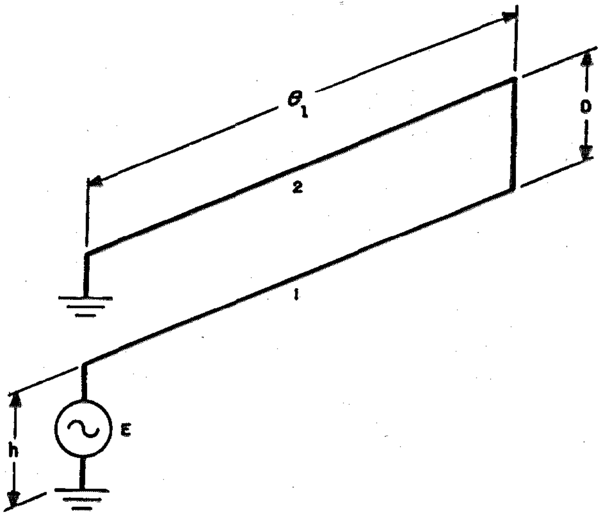

Fig. 2.

This has been previously analyzed by an artificial arrangement of exciting generators shown in Fig. 3. This arrangement clearly excites line *1* at potential E and line *2* at potential zero. The current in line *1* may be determined by adding the effects of (1) the balanced generator pair and (2) the generator exciting the line against ground.

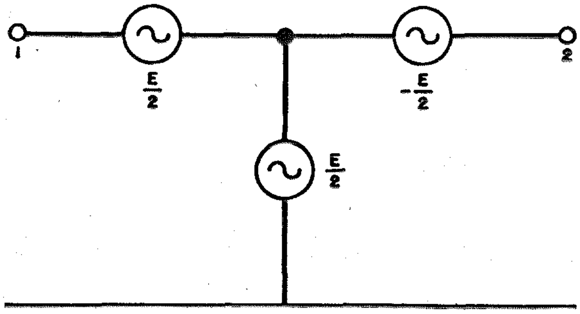

Fig. 3.

For part (1), the line is a shorted line of length θ_l; its input admittance is

$$Y_b = -iY_{ob} \cot \theta_l,$$

and the balance current into conductor *1* is

$$I_b = -iY_{ob}E \cot \theta_l. \qquad (22A)$$

For part (2), the line is an open line of length θ_l; the input admittance to conductor *1* is

$$Y_u = i\left(\frac{Y_{ou}}{2}\right) \tan \theta_l,$$

and the unbalance current into conductor *1* is

$$I_u = i\left(\frac{Y_{ou}}{2}\right)\left(\frac{E}{2}\right) \tan \theta_l. \qquad (22B)$$

The input admittance at conductor *1* is, therefore,

$$Y_1 = \frac{I_b + I_u}{E} = i(\tfrac{1}{4}Y_{ou} \tan \theta_l - Y_{ob} \cot \theta_l). \qquad (23)$$

Equation (23) is obtainable from the foregoing theory in a straightforward manner. In (9), with $n = 2$, we introduce the following boundary conditions (Fig. 2):

$$\begin{aligned}V_1 &= E, &&\text{when } \theta = 0,\\ V_1 &= V_2, &&\text{when } \theta = \theta_l,\\ V_2 &= 0, &&\text{when } \theta = 0,\\ I_1 &= -I_2, &&\text{when } \theta = \theta_l.\end{aligned}$$

Setting $s = e\ \exp[i\theta_l]$ yields a set of equations:

$$\begin{aligned}&A_1 + B_1 = E,\\ &s^{-1}A_1 + sB_1 = s^{-1}A_2 + sB_2,\\ &A_2 + B_2 = 0,\\ &s^{-1}A_1 - sB_1 = -s^{-1}A_2 + sB_2,\end{aligned}$$

which can be solved for the A's and B's. The input current to line *1* in terms of E is then completely known; in fact this current is

$$(I_i)_0 = Y_{11}(A_1 - B_1) + Y_{12}(A_2 - B_2),$$

whence the admittance is found to agree with (23).

3. *"Loop" or "Short-to-Short" Coupling Between Lines*

The configuration to be analyzed is shown in Fig. 4. Line *1* is driven and power is induced in line *2* and absorbed in the load $2Y_0$. The problem is to find the input admittance at $M-M$ on line *1*. The solution in the present instance is restricted to the case where the geometry of the lines is identical and where the transverse traces of the centers of the wires constitute the corners of a rectangle.

This problem has been studied previously;[1] in the earlier discussion, the characteristic impedances of the driving and driven lines were considered to be unaffected by their proximity to each other. The effect of proximity is accounted for automatically in the present procedure.

To fit the problem into the present discussion, consider both lines balanced with respect to ground so that the configuration of Fig. 1 can be substituted. Then the boundary conditions are

$$\begin{aligned} &V_1 = E, && \text{when } \theta = 0, \\ &V_1 = 0, && \text{when } \theta = \theta_l, \\ &V_2 = 0, && \text{when } \theta = 0, \\ &I_2 = 2Y_0(V_2)_{\theta-\theta_l}, && \text{when } \theta = \theta_l. \end{aligned}$$

The current in line *1* at $\theta = 0$ is

$$(I_i)_0 = Y_{11}(A_1 - B_1) + Y_{12}(A_2 - B_2), \tag{24}$$

and the input admittance is

$$Y_{in} = \frac{(I_i)_0}{2E}. \tag{25}$$

Substitution of the boundary conditions in (8) and (9) yields the following set of equations.

$$\begin{gathered} A_1 + B_1 = E, \\ s^{-1}A_1 + sB_1 = 0, \\ A_2 + B_2 = 0, \end{gathered}$$

$$s^{-1}Y_{21}A_1 - sY_{21}B_1 + s^{-1}(Y_{22} - 2Y_0)A_2 - s(Y_{22} + 2Y_0)B_2 = 0.$$

These equations are readily solved to yield

$$\left.\begin{aligned} A_1 &= \frac{-Es}{s - s^{-1}}, \\ B_1 &= -\frac{Es^{-1}}{s - s^{-1}}, \\ A_2 &= -\frac{2Y_{12}E}{(s - s^{-1})[(s + s^{-1})Y_{22} + 2(s - s^{-1})Y_0]}, \\ B_2 &= -A_2. \end{aligned}\right\} \tag{26}$$

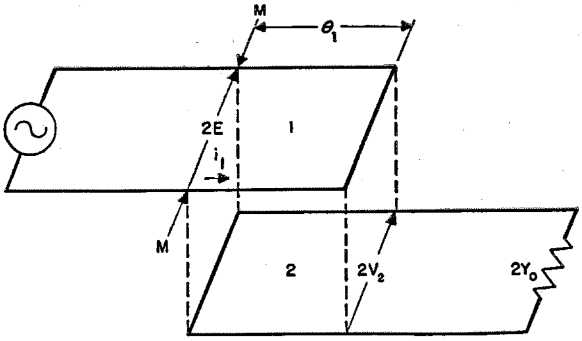

Fig. 4.

Substitution of (26) in (24) and (25) gives, after simple reductions,

$$Y_{in} = \frac{Y_{11}}{2i}\cot\theta_1 - \frac{Y_{12}^2}{2i\sin\theta_1(Y_{22}\cos\theta_1 + i2Y_0\sin\theta_1)}$$

or, finally,

$$Y_{in} = \frac{Y_{11}}{2i}\cot\theta_1\left[1 - \frac{Y_{12}^2}{Y_{11}} \times \frac{1}{\cos\theta_1(Y_{11}\cos\theta_1 + i2Y_0\sin\theta_1)}\right] \tag{27}$$

as $Y_{22} = Y_{11}$ for the configuration discussed.

This result is found to check the earlier result[1] when the separation between the two lines is made sufficiently great.

Of course, it must be pointed out that the simple procedures given here make no allowances for end effects on the lines; such effects are minimized when the spacing between conductors is small compared with line length.

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[1] A. Alford, "Coupled Networks in Radio-Frequency Circuits," *Proceedings of the I.R.E.*, v. 29, pp. 55–69; February, 1941: see p. 68.

Reduction by Limiters of Amplitude Modulation in an Amplitude- and Frequency-Modulated Wave

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A MATHEMATICAL method is described for computing the effect of idealized limiters on an amplitude- and frequency-modulated wave of the form $A \sin \phi(t)$, where A may be a function of time. The output components of the limiter are shown to be of the form $V_{2\nu+1}(A) \sin (2\nu+1)\phi(t)$ and the dynamic characteristics $V_{2\nu+1}(A)$ are used to compute the residual amplitude modulation in the output. The dynamic characteristics are plotted for several limiters assumed to behave according to known mathematical functions. The effectiveness of such limiters is displayed in curves showing the ratio of residual amplitude-modulation index to the amplitude-modulation index in the original wave as a function of the applied voltage.

• • •

1. *Introduction*

In many problems, the action of a four-terminal network is characterized by a curve showing the output V in terms of the input v. This curve may not be linear and such a case may be treated in a number of ways. One approach is to represent the output-versus-input characteristic by a power series in the form

$$V = f(v) = \sum_{n=0}^{\infty} a_n v^n. \tag{1}$$

Another approach is to represent the characteristic function by means of a trigonometric series[1] in the form

$$V = A_0 + \sum_{1}^{\infty} \left(A_n \cos \frac{2\pi n v}{v_0} + B_n \sin \frac{2\pi n v}{v_0} \right), \tag{2}$$

where v_0 equals the range over which this expansion is valid.

The representations given in (1) and (2) are useful in numerical analysis where only a few terms are required to give an approximate result. For general analysis, (1) is applicable only to continuous functions and is very difficult to handle as the result is in the form of an infinite series. Equation (2), while giving rise to an infinite series, may sometimes be treated by the method given in the appendix.

A third approach, which is applied in this paper, makes use of the Fourier transform. It is similar to the contour-integral method used by Bennett[2] and others but has the advantage that the integrals involved are real.

2. *Outline of Method*

The specific problem that will be treated here is the action of amplitude limiters on a signal that may be modulated both in amplitude and frequency. Our discussion will be limited to the case where the static characteristic is an odd function of the input voltage v, so that

$$f(v) = -f(-v),$$

where $v = A(t) \sin \phi(t)$, $A(t)$ and $\phi(t)$ being given functions of time.

To compute the output of the limiter, use will be made of the Fourier transform in the following manner. Any limiter characteristic may be represented approximately by Fig. 1*A*.

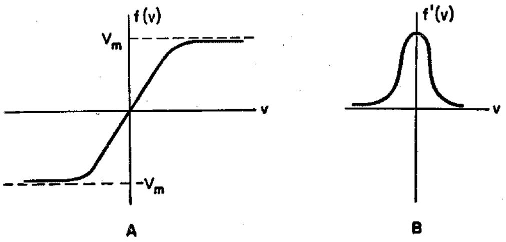

Fig. 1—*A* is the static characteristic of the limiter, and *B* is the derivative of that static characteristic.

While there are a number of known functions that may be used to realize such a characteristic, there is obviously one common property to all of them, namely: that the derivative $f'(v)$ of the limiter curve is finite at $v=0$ and tends to zero as $v \rightarrow \pm\infty$, also $f'(v)=f'(-v)$ as shown in Fig. 1*B*. This property of the derivative $f'(v)$ enables its Fourier transform to be found without difficulty. Following the notation used in a previous paper,[3] let

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[1] L. W. Barrow, "Contribution to the Theory of Non-linear Circuits with Large Applied Voltages," *Proceedings of the I.R.E.*, v. 22, pp. 964–980; August, 1934.

[2] W. R. Bennett, "Response of a Linear Rectifier to Signal Noise," *Bell System Technical Journal*, v. 23, pp. 97–113; January, 1944.

$$\mathcal{T}_v^u f'(v)=g(u)=\frac{2}{(2\pi)^{\frac{1}{2}}}\int_0^\infty f'(v)\cos(uv)dv. \quad (3)$$

In the particular case considered for which $f(v)$ is bounded everywhere, it can be shown that the Fourier transform of $f(v)$ is related to the Fourier transforms of $f'(v)$ by

$$\mathcal{T}_v^u f(v)=\frac{1}{ju}\mathcal{T}_v^u f'(v).$$

$$\therefore \quad G(u)=\mathcal{T}_v^u f(v)=\frac{1}{ju}\mathcal{T}_v^u f'(v)$$

$$=\frac{2}{ju(2\pi)^{\frac{1}{2}}}\int_0^\infty f'(v)\cos(uv)dv.$$

The output signal of the limiter is given by

$$V=f(v)=\mathcal{T}_u^v G(u)=\frac{2j}{(2\pi)^{\frac{1}{2}}}\int_0^\infty G(u)\sin(uv)du$$

$$=\frac{2}{\pi}\int_0^\infty\frac{\sin(uv)du}{u}\int_0^\infty f'(v)\cos(uv)dv. \quad (4)$$

Here V is a function of the input voltage v expressed in the form of a definite integral. Since $v=A(t)\sin\phi(t)$ and $\sin(uv)=2\sum_{\nu=0}^{\infty}J_{2\nu+1}(uA)$ $\times\sin(2\nu+1)\phi(t)$, the coefficient of $\sin(2\nu+1)\phi(t)$ in the output may be defined by analogy with the case of a simple sine wave as the amplitude of the $(2\nu+1)$th harmonic of the fundamental and is given by

$$V_{2\nu+1}(A)=\frac{4}{\pi}\int_0^\infty\frac{J_{2\nu+1}(uA)du}{u}\int_0^\infty f'(v)\cos(uv)dv. \quad (5)$$

In particular, the amplitude of the fundamental voltage is of interest.

$$V_1(A)=\frac{4}{\pi}\int_0^\infty\frac{J_1(uA)du}{u}\int_0^\infty f'(v)\cos(uv)dv. \quad (6)$$

In practical cases where the limiter is followed by a frequency discriminator for demodulation, it is obvious that the frequency of the frequency-modulated carrier should be chosen so that the significant sidebands of the harmonics should not fall within the sidebands of the fundamental.

3. Application to Several Functions that Approximate an Amplitude Limiter

Following the general idea outlined above, limiter action may be analyzed by curves, the closest to the experimental result being finally chosen. Several known mathematical functions, which represent such curves and their derivatives, are given in Figs. 2–6.

The effectiveness of the various limiters listed above is clearly shown when their derivatives are compared. While the derivative of the semi-ideal limiter drops suddenly to zero, the derivative of the inverse-tangent limiter varies as $1/(1+x^2)$. However, the derivative of the error-integral limiter varies as e^{-x^2}, which indicates a high degree of limiting.

In the following, the expression given for $f'(v)$ for the various limiter curves outlined above will be substituted in (4) from which the dynamic characteristics and the reduction of amplitude modulation will be evaluated.

3.1 IDEAL AND SEMI-IDEAL LIMITERS

Consider the semi-ideal limiter $f(v)$ shown in Fig. 2*B* and its derivative $f'(v)$ in Fig. 2*A*. Starting with $f'(v)$ and applying (4), we have

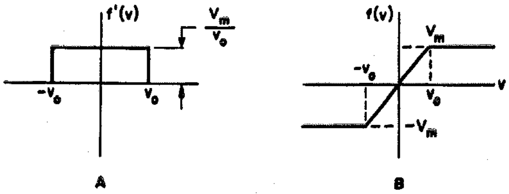

Fig. 2—*A* is the derivative of a semi-ideal limiter, and *B* is its static characteristic.

$$\int_0^\infty f'(v)\cos(uv)dv=\frac{V_m}{v_0}\int_0^{v_0}\cos(uv)dv$$

$$=\frac{V_m}{v_0}\frac{\sin(uv_0)}{u}.$$

$$\therefore \quad f(v)=\frac{2}{\pi}\frac{V_m}{v_0}\int_0^\infty\frac{\sin(uv_0)\sin(uv)du}{u^2}, \quad (7)$$

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[3] A. G. Clavier, P. F. Panter, and D. D. Grieg, "PCM Distortion Analysis," *Electrical Engineering*, v. 66, pp. 1110–1122; November, 1947.

when $v_0 \to 0$, (7) reduces to

$$f(v) = \frac{2}{\pi} V_m \int_0^\infty \frac{\sin (uv) du}{u}, \tag{8}$$

which is the equation of the ideal limiter as shown in Fig. 3B. The dynamic characteristic of the $(2\nu+1)$th harmonic is given by[4]

$$\left.\begin{aligned} V_{2\nu+1}(A) &= \frac{4 V_m}{\pi} \int_0^\infty \frac{J_{2\nu+1}(uA) du}{u} \\ &= \frac{4 V_m}{\pi} \frac{1}{2\nu+1}. \end{aligned}\right\} \tag{9}$$

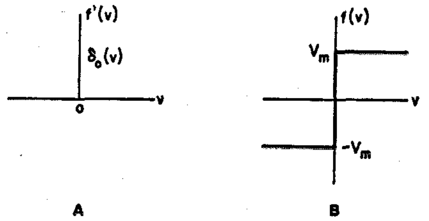

Fig. 3—*A* is the derivative of an ideal limiter, and *B* is its static characteristic.

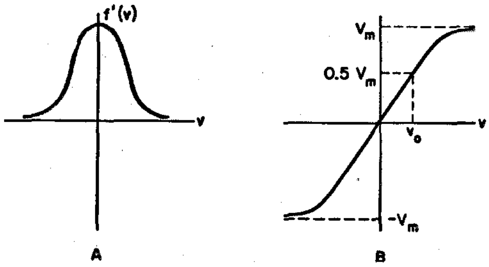

Fig. 4—*A* is the derivative of an inverse-tangent limiter, $f'(v) = \frac{2}{\pi} \frac{V_m}{v_0} \frac{v_0^2}{v^2+v_0^2}$. *B* is the static characteristic, $f(v) = \frac{2}{\pi} \times V_m \tan^{-1} \frac{v}{v_0}$.

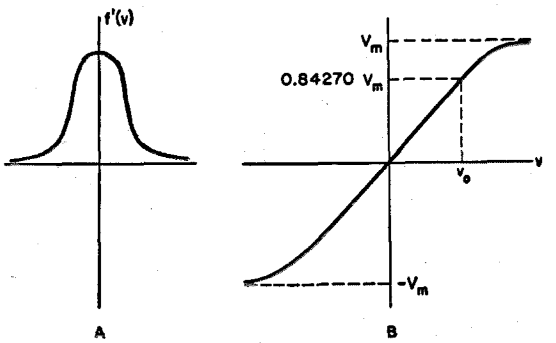

Fig. 5—*A* is the derivative of an error-integral limiter, $f'(v) = \frac{V_m}{v_0} \frac{2}{\pi^{\frac{1}{2}}} e^{-(v/v_0)^2}$. *B* is the static characteristic, $f(v) = \frac{2 V_m}{\pi^{\frac{1}{2}}} \times v_0 \int_0^v e^{-(v/v_0)^2} dv$.

It is obvious from (9) that the output signal of an ideal limiter is independent of the amplitude of the input. It is also apparent that the ideal limiter may be used as a frequency multiplier.

To continue with the semi-ideal limiter, from (5) may be obtained an expression for its dynamic characteristic of the $(2\nu+1)$th harmonic.

$$V_{2\nu+1}(A) = \frac{4 V_m}{\pi v_0} \int^\infty \frac{J_{2\nu+1}(uA) \sin (uv_0) du}{u^2}.$$

This integral may be evaluated by the relation

$$J_{n+1}(x) + J_{n-1}(x) = \frac{2n}{x} J_n(x)$$

and (2) of reference 4, page 405.

$$\therefore \quad V_{2\nu+1}(A) = \frac{2 V_m A}{\pi v_0 (2\nu+1)} \times \int_0^\infty \frac{J_{2\nu}(uA) + J_{2\nu+2}(uA)}{u} \sin (uv_0) du.$$

The fundamental is given by

$$\begin{aligned} V_1(A) &= A\left(\frac{V_m}{v_0}\right), \quad A \leqslant v_0, \\ &= \frac{2}{\pi} A \left(\frac{V_m}{v_0}\right) \left\{ \sin^{-1} \frac{v_0}{A} + \frac{v_0}{A} \left[1 - \left(\frac{v_0}{A}\right)^2 \right]^{\frac{1}{2}} \right\}, \end{aligned} \tag{10}$$

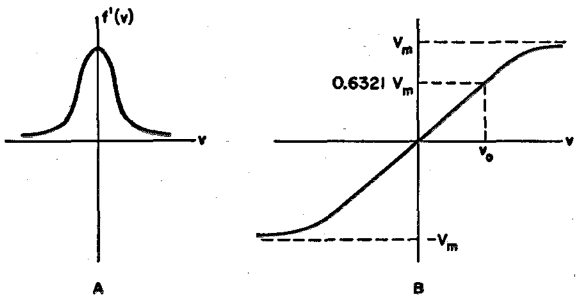

Fig. 6—*A* is the derivative of an exponential limiter, $f'(v) = \frac{V_m}{v_0} e^{-(v_0/v)}$, $v \geqslant 0$; $f'(v) = \frac{V_m}{v_0} e^{(v/v_0)}$, $v \leqslant 0$. *B* is the static characteristic, $f(v) = V_m(1 - e^{-(v/v_0)})$, $v \geqslant 0$; $f(v) = -V_m(1 - e^{(v/v_0)})$, $v \leqslant 0$.

---

[4] G. N. Watson, "A Treatise on the Theory of Bessel Functions," 2nd Edition, Cambridge University Press, Cambridge, England, 1944. Page 391, equation 1.

$A \geqslant v_0$, and the harmonics are given by

$$V_{2\nu+1}(A)=0, \quad A \leqslant v_0,$$

$$=\frac{A}{\pi(2\nu+1)}\left(\frac{V_m}{v_0}\right)\left\{\frac{\sin\left[2\nu \sin^{-1}\frac{v_0}{A}\right]}{\nu}+\frac{\sin\left[(2\nu+2)\sin^{-1}\frac{v_0}{A}\right]}{\nu+1}\right\}, \quad A \geqslant v_0. \quad (11)$$

Thus we see from (10) and (11) that on the linear portion of the static curve where $A \leqslant v_0$, the output is proportional to the input, as expected, and the harmonics are equal to zero. However, in the range $A > v_0$, harmonics are generated as a result of the nonlinearity of the system, the degree of limiting of the fundamental and its harmonics depending on the amplitude A of the input signal. When $A \gg v_0$, (10) of the semi-ideal limiter reduces to (9) of the ideal one, where the limiting is perfect. The dynamic characteristics of the semi-ideal limiter are shown in Fig. 7.

3.2 INVERSE-TANGENT LIMITER

Referring to Figs. 4*A* and 4*B*, the derivative of the inverse-tangent limiter is given by

$$\therefore \quad f'(v)=\frac{2}{\pi}\left(\frac{V_m}{v_0}\right)\frac{v_0^2}{u^2+v_0^2},$$

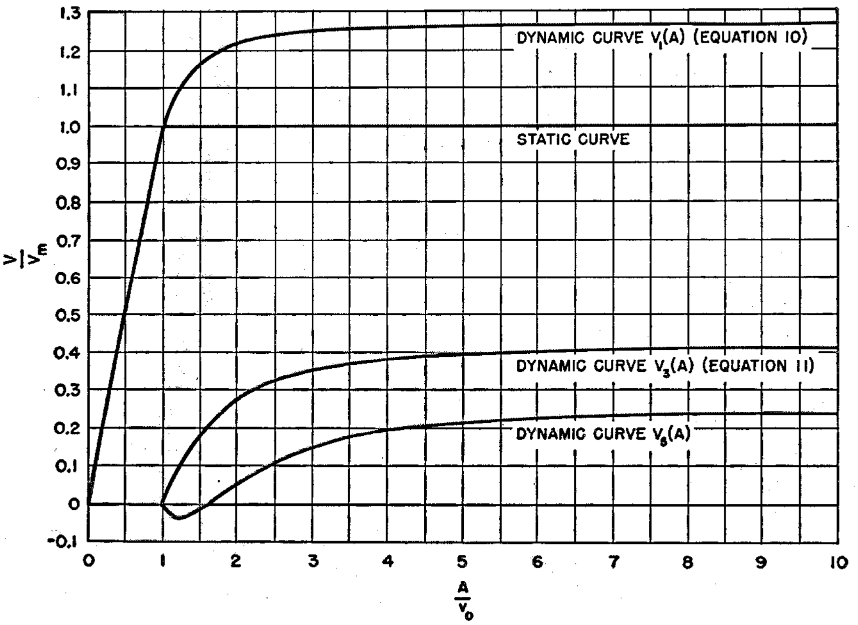

Fig. 7—Semi-ideal limiter.

and its static curve is given by

$$F(v)=\frac{2}{\pi}V_m \tan^{-1}\frac{v}{v_0}. \quad (12)$$

Following the procedure outlined above, we have

$$\int_0^\infty f'(v)\cos(uv)dv=\frac{2V_m v_0}{\pi}\int_0^\infty \frac{\cos(uv)}{u^2+v_0^2}dv$$

$$=V_m e^{-uv_0}.$$

$$\therefore \quad f(v)=\frac{2V_m}{\pi}\int_0^\infty \frac{e^{-uv_0}\sin(uv)du}{u} \quad (13)$$

and the dynamic characteristics are given by[5]

$$V_{2\nu+1}(A)=\frac{4V_m}{\pi}\int_0^\infty \frac{e^{-uv_0}J_{2\nu+1}(uA)du}{u}$$

$$=\frac{4V_m}{\pi(2\nu+1)}\left\{\left[1+\left(\frac{v_0}{A}\right)^2\right]^{\frac{1}{2}}-\frac{v_0}{A}\right\}^{2\nu+1}. \quad (14)$$

The static curve and its dynamic characteristics are plotted in Fig. 8. As expected, such a limiter is less desirable than the semi-ideal one.

3.3 LIMITER EQUATION GIVEN BY ERROR INTEGRAL

The static curve and its derivative are shown in Fig. 5, then

$$\left.\begin{aligned} f(v)&=\frac{2V_m}{v_0\pi^{\frac{1}{2}}}\int_0^\infty e^{-(v/v_0)^2}dv,\\ f'(v)&=\frac{V_m}{v_0}\frac{2}{\pi^{\frac{1}{2}}}e^{-(v/v_0)^2}.\end{aligned}\right\} \quad (15)$$

$$\therefore \quad \int_0^\infty f'(v)\cos(uv)dv$$

$$=\frac{2V_m}{v_0\pi^{\frac{1}{2}}}\int_0^\infty e^{-(v/v_0)^2}\cos(uv)dv$$

$$=V_m e^{-(uv_0/2)^2}.$$

$$\therefore \quad f(v)=\frac{2V_m}{\pi}$$

$$\times\int_0^\infty \frac{e^{-(uv_0/2)^2}\sin(uv)du}{u}. \quad (16)$$

---

[5] Reference 4, page 386, equation 7.

The dynamic characteristics are given by[6]

$$
\begin{aligned}
V_{2\nu+1}(A) &= \frac{4V_m}{\pi}\int_0^\infty \frac{e^{-(uv_0/2)^2}J_{2\nu+1}(uA)du}{u} \\
&= \frac{2V_mA}{\pi(2\nu+1)}\int_0^\infty e^{-(uv_0/2)^2} \\
&\qquad \times[J_{2\nu}(uA)+J_{2\nu+2}(uA)]du \\
&= \frac{2V_m}{\pi^{\frac{1}{2}}}\frac{A}{v_0}\frac{e^{-(A^2/2v_0^2)}}{(2\nu+1)} \\
&\qquad \times\left[I_\nu\left(\frac{A^2}{2v_0^2}\right)+I_{\nu+1}\left(\frac{A^2}{2v_0^2}\right)\right], \quad (17)
\end{aligned}
$$

where $I_\nu(x)$ is the modified Bessel function of order ν. The dynamic characteristic of the fundamental and its harmonics is shown in Fig. 9, making use of available tables of $e^{-x}I_\nu(x)$ and $e^{-x}I_{\nu+1}(x)$ as given in reference 4.

3.4 LIMITER EQUATION GIVEN BY AN EXPONENTIAL

This is shown in Fig. 6.

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[6] Reference 4, page 394, equation 5.

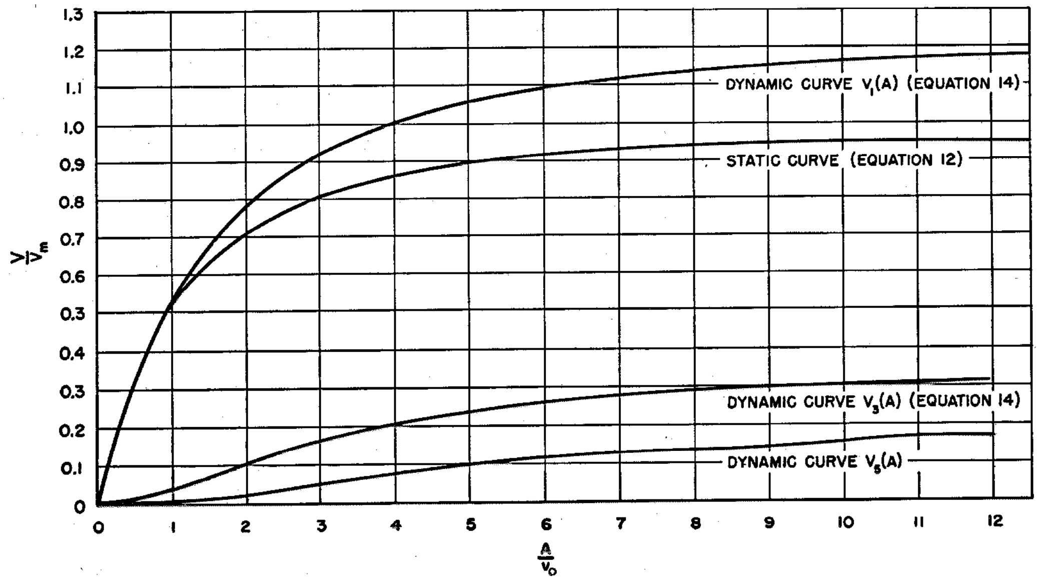

Fig. 8—Inverse-tangent limiter.

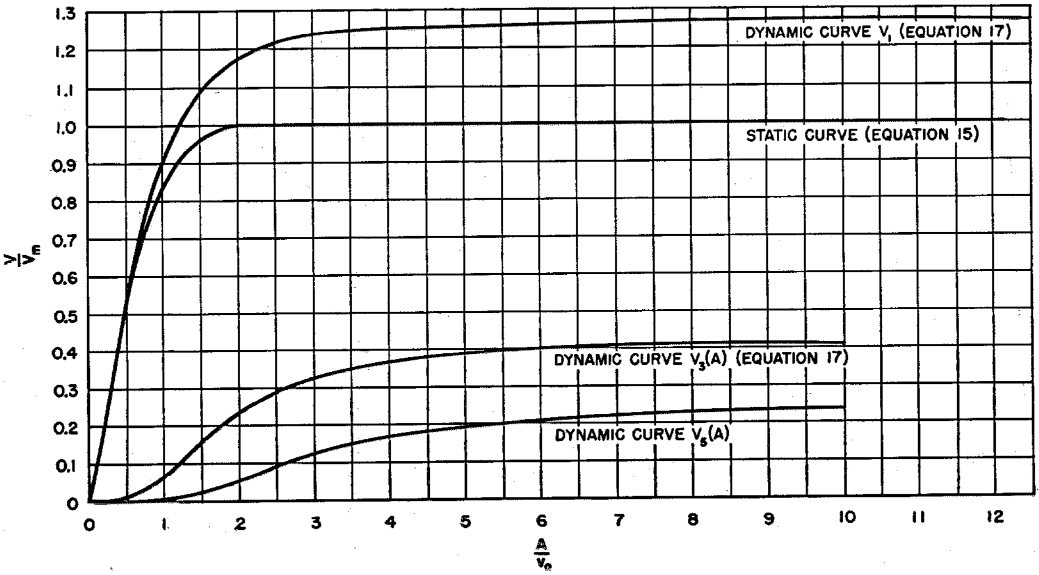

Fig. 9—Error-integral limiter.

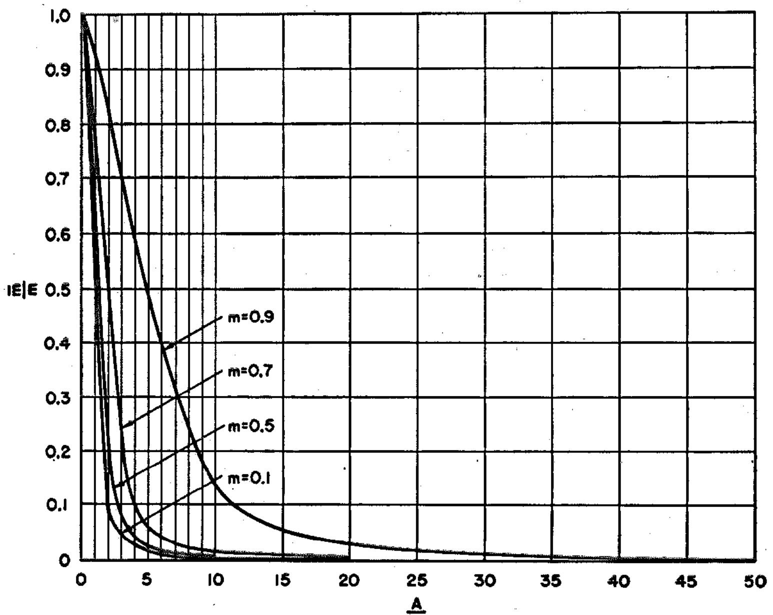

Fig. 10—Reduction in amplitude modulation versus input voltage for a semi-ideal limiter.

$$
\left.
\begin{aligned}
f(v) &= V_m(1-e^{-(v/v_0)}), & v \geqslant 0,\\
&= -V_m(1-e^{v/v_0}), & v \leqslant 0,\\
\text{and}\quad f'(v) &= \frac{V_m}{v_0}\, e^{-(v/v_0)}, & v \geqslant 0,\\
&= \frac{V_m}{v_0} e^{v/v_0}, & v \leqslant 0.
\end{aligned}
\right\} \qquad (18)
$$

$$
\therefore \quad \int_0^\infty f'(v) \cos(uv)dv
$$

$$
= \frac{V_m}{v_0} \int_0^\infty e^{-(v/v_0)} \cos(uv)dv
$$

$$
= V_m \frac{1}{1+(uv_0)^2},
$$

$$
f(v) = \frac{2V_m}{\pi} \int_0^\infty \frac{\sin(uv)du}{u(1+\overline{uv}_0^2)}. \qquad (19)
$$

The dynamic characteristics are given by

$$
V_{2\nu+1}(A) = \frac{4V_m}{\pi} \int_0^\infty \frac{J_{2\nu+1}(uA)du}{u(1+\overline{uv}_0^2)}. \qquad (20)
$$

The fundamental reduces to[7]

$$
\begin{aligned}
V_1(A) &= \frac{4V_m}{\pi} \int_0^\infty \frac{J_1(uA)du}{u(1+\overline{uv}_0^2)}\\
&= 2V_m\left[I_1\left(\frac{A}{v_0}\right) - L_1\left(\frac{A}{v_0}\right)\right]. \qquad (21)
\end{aligned}
$$

Where the function $L_\nu(x)$ is defined by the equation[8]

$$
L_\nu(x) = \sum_{m=0}^{\infty} \frac{(x/2)^{\nu+2m+1}}{\Gamma(m+\frac{3}{2})\Gamma(\nu+m+\frac{3}{2})},
$$

the function $L_\nu(x)$ bears the same relation to Struve's function as $I_\nu(x)$ bears to $J_\nu(x)$.

No curves were plotted for this limiter due to the fact that the authors are not aware of tables listing $L_1(x)$.

4. *Reduction of Amplitude Modulation*

Let $A(t)$ be an amplitude-modulated wave, namely, $A(t) = A_0(1+m \sin pt)$. The amplitude of the input varies, therefore, between the limits $A_0(1 \pm m)$. The effect of the limiter on such a wave is to reduce the index of modulation m. The output index of amplitude modulation will be defined by the relation

$$
\bar{m} = \frac{V_1[A_0(1+m)] - V_1[A_0(1-m)]}{V_1[A_0(1+m)] + V_1[A_0(1-m)]}, \qquad (22)
$$

which may be obtained from the dynamic curve of the fundamental. The ratio $\bar{m}/m$, when plotted versus the input voltage v, gives a measure of the reduction in amplitude modulation effected by the action of the limiter.

Curves showing the variation of $\bar{m}/m$ versus A/v_0 are plotted in Figs. 10–12, for the limiters discussed above. The degree of limiting as a function of the input voltage may also be investigated by expanding $V_1(A)$ in powers of v_0/A. For the semi-ideal limiter,

$$
\begin{aligned}
V_1(A) &= \frac{2V_m}{\pi}\left\{\frac{A}{v_0} \sin^{-1} \frac{v_0}{A} + \left[1-\left(\frac{v_0}{A}\right)^2\right]^{\frac{1}{2}}\right\}\\
&= \frac{4V_m}{\pi}\left[1-\frac{1}{6}\left(\frac{v_0}{A}\right)^2 - \frac{1}{40}\left(\frac{v_0}{A}\right)^4\right]. \qquad (23)
\end{aligned}
$$

For the error-integral limiter, using the asymptotic expansion of $I_0(x)$ and $I_1(x)$,

$$
V_1(A) = \frac{4V_m}{\pi}\left[1-\frac{1}{4}\left(\frac{v_0}{A}\right)^2 - \frac{3}{32}\left(\frac{v_0}{A}\right)^4 \cdots\right].
$$

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[7] Reference 4, page 426, equation 11.

[8] Reference 4, page 329, equation 11.

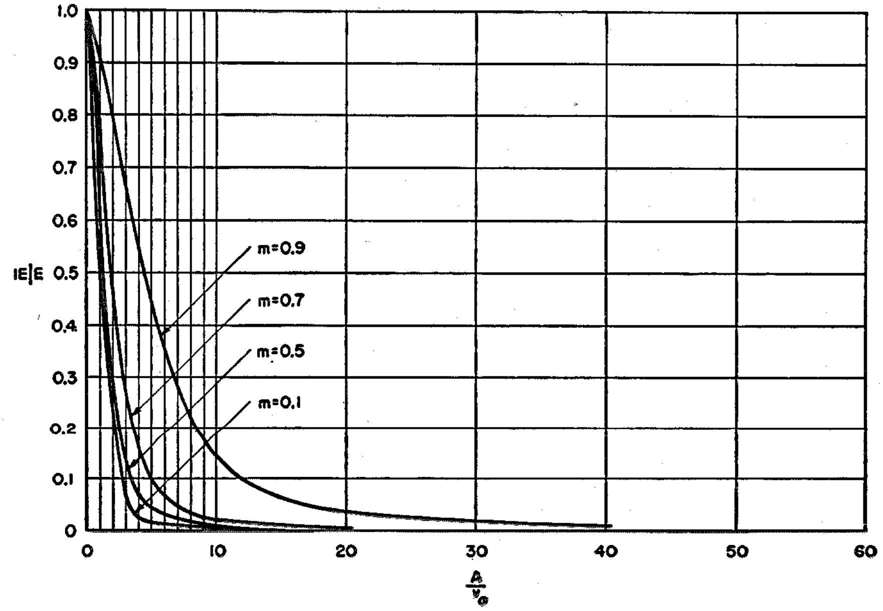

Fig. 11—Reduction in amplitude modulation versus input voltage for an error-integral limiter.

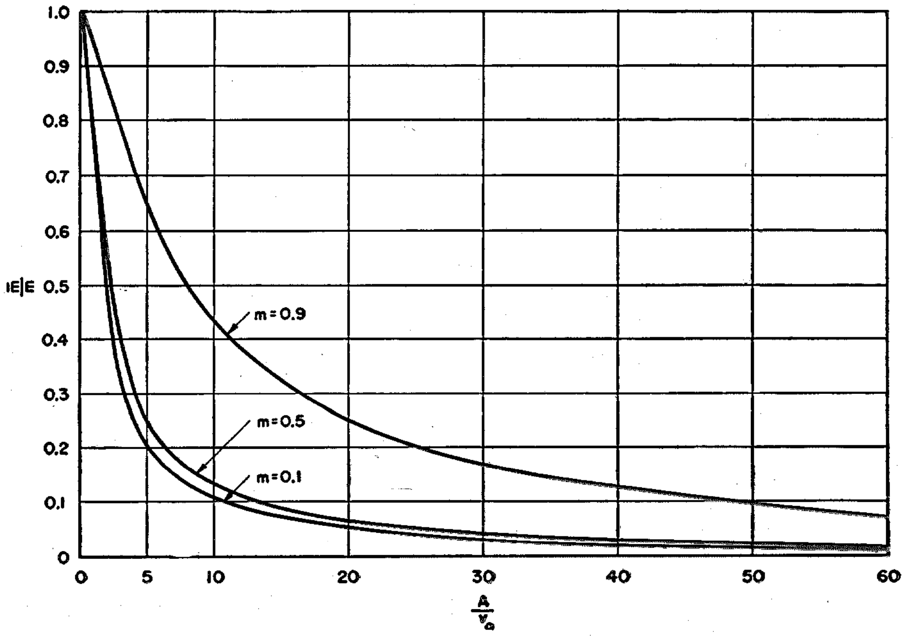

Fig. 12—Reduction in amplitude modulation versus input voltage for an inverse-tangent limiter.

Using these relations and (22) for $\bar{m}/m$, for the semi-ideal limiter

$$\frac{\bar{m}}{m}=\frac{1}{3}\left(\frac{v_0}{A_0}\right)\frac{1}{(1-m^2)^2}, \qquad (24)$$

and for the error-integral limiter

$$\frac{\bar{m}}{m}=\frac{1}{2}\left(\frac{v_0}{A_0}\right)^2\frac{1}{(1-m^2)^2}. \qquad (25)$$

The last two equations illustrate the effectiveness of these limiters, which is also borne out in Figs. 10 and 11. However, for the inverse-tangent limiter,

$$V_1(A)=\frac{4V_m}{\pi}\left\{\left[1+\left(\frac{v_0}{A}\right)^2\right]^{\frac{1}{2}}-\frac{v_0}{A}\right\}$$
$$=\frac{4V_m}{\pi}\left[1-\frac{v_0}{A}+\frac{1}{2}\left(\frac{v_0}{A}\right)^2\cdots\cdot\right].$$

In this case

$$\frac{\bar{m}}{m}\doteq\frac{v_0}{A_0}\frac{1}{1-m^2}. \tag{26}$$

The performance of such a limiter in the reduction of the index of modulation would then be much poorer than the preceding ones, which is clearly illustrated in Fig. 12.

5. *Conclusions*

It has been shown that the effectiveness of a limiter is conveniently described by means of the derivative of the static characteristic curve. A number of different equations have been chosen for this derivative and the effect of the corresponding limiters on a wave having both frequency and amplitude modulation has been evaluated by means of Fourier transforms. The output has been found to consist of a series of terms of the same type as the input wave, but the phase functions of which are multiplied by odd integers. These terms have been called fundamental and harmonics. In general, the carrier frequency of the input wave is chosen so that the so-called harmonics can in practice be filtered out from the fundamental. The fundamental presents a residual amplitude modulation and the ratio of this residual amplitude-modulation index to the output amplitude-modulation index has been plotted versus the input voltage for the different cases of limiter characteristics that have been examined.

The method used can obviously be extended to other nonlinear four-terminal networks including, for instance, rectifiers or vacuum tubes, when the effect of saturation is considered.

6. *Appendix*

The expressions given for the semi-ideal limiter shown in Fig. 13 may be obtained by expanding the static curve as a Fourier series valid over the range $-V_1$ to $+V_1$.

The characteristic is defined by the equations:

$$\left.\begin{aligned} V&=-V_m, & -v_1&\leqslant v\leqslant -v_0.\\ V&=\frac{V_m}{v_0}v, & v_0&\leqslant v\leqslant v_0.\\ V&=+V_m, & v_0&\leqslant v\leqslant v_1.\end{aligned}\right\} \tag{27}$$

To simplify the results, we may subtract the linear function defined by

$$V'=\frac{V_m}{v_1}v, \quad -v_1\leqslant v\leqslant v_1. \tag{28}$$

This gives the function shown in Fig. 14 defined by

$$\left.\begin{aligned} V''&=-V_m-\frac{V_m}{v_1}v, & -v_1&\leqslant v\leqslant -v_0.\\ V''&=v\left(\frac{V_m}{v_0}-\frac{V_m}{v_1}\right), & -v_0&\leqslant v\leqslant v_0.\\ V''&=V_m-\frac{V_m}{v_1}v, & v_1&\leqslant v\leqslant v_0.\end{aligned}\right\} \tag{29}$$

The function V'' is an odd function so that it may be expanded into a sine series as follows.

$$f(v)=\sum_{n=1}^{\infty}a_n\sin\frac{2\pi nv}{v_1}, \tag{30}$$

where

$$a_n=\frac{2}{v_1}\int_0^{v_1}f(v)\sin\frac{2\pi nv}{v_1}dv.$$

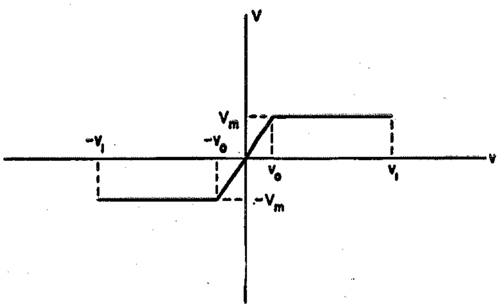

Fig. 13—Semi-ideal limiter.

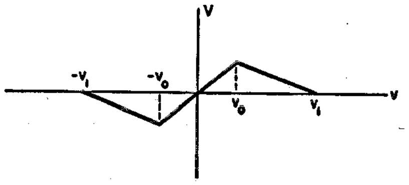

Fig. 14—The function V''.

Performing the integrations, we have as the equation of the static limiter curve

$$V=\frac{2V_mv}{v_1}+\frac{4V_mv_0}{v_1}\sum_{n=1}^{\infty}\frac{\sin\dfrac{2\pi nv_0}{v_1}}{\left(\dfrac{2\pi nv_0}{v_1}\right)^2}\sin\frac{2\pi nv}{v_1}$$

$$=-\frac{2V_mv}{v_1}+\frac{4V_mv_0}{v_1}\sum_{0}^{\infty}\frac{\sin\dfrac{2\pi nv_0}{v_1}}{\left(\dfrac{2\pi nv_0}{v_1}\right)^2}\sin\frac{2\pi nv}{v_1}. \qquad (31)$$

If we let $v=A\sin\phi$, the coefficient of $\sin\phi$ is

$$V_1(A)=-\frac{2V_m}{v_1}A+\frac{8V_mv_0}{v_1}\sum_{0}^{\infty}\frac{\sin\dfrac{2\pi nv_0}{v_1}}{\left(\dfrac{2\pi nv_0}{v_1}\right)^2}J_1\left(\frac{2\pi nA}{v_1}\right). \qquad (32)$$

To sum the infinite series, we may use the Euler-Maclaurin summation formula

$$f(a)+f(a+w)+\ldots.+f[a+(n-1)w]+f(a+nw)$$
$$=\frac{1}{w}\int_a^{a+nw}f(x)dx+\tfrac{1}{2}[f(a)+f(a+nw)]$$
$$-\frac{B_1w}{2!}[f'(a+nw)-f'(a)]$$
$$+\frac{B_2w^3}{4!}[f'''(a+nw)-f'''(a)]\ldots., \qquad (33)$$

where $B_1, B_2, \ldots.$ are Bernoulli's numbers. Here $a=0$, $w=2\pi/v_1$, so that

$$f(x)=\frac{\sin v_0x}{v_0^2x^2}J_1(Ax). \qquad (34)$$

As $f(x)$ is an even function, all the odd derivatives vanish when $x = 0$. As x is allowed to become infinite, the function and its derivatives vanish. Thus:

$$\sum_{0}^{\infty}\frac{\sin\dfrac{2\pi nv_0}{v_1}J_1\left(\dfrac{2\pi nA}{v_1}\right)}{\left(\dfrac{2\pi nvv_0}{v}\right)^2}$$
$$=\frac{v_1}{2\pi v_0^2}\int_0^{\infty}\frac{\sin v_0xJ_1(Ax)dx}{x^2}+\frac{1}{4}\frac{A}{v_0}.$$

This gives

$$V_1(A)=\frac{4V_m}{v_0}\int_0^{\infty}\frac{\sin v_0xJ_1(Ax)dx}{x^2}, \qquad (35)$$

which is the same formula as derived previously. The result is independent of the assumed range $-v_1$ to v_1 and hence is a general result. The expressions for the higher-order characteristics may be obtained in the same manner.

Thermomagnetic Generator

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PROPOSALS have been made[1] to use the properties of ferromagnetic substances near the Curie point for constructing thermomagnetic current generators. Large flux variations could be induced in a coil and alternating currents generated by heating and cooling above and below its Curie point a ferromagnetic substance that is part of a magnetic circuit.

The mechanism, therefore, lies in the temperature dependence of the permeability of the substance and the associated thermomagnetic effects. In this paper, calculations are given of the magnitude of the effect and of the efficiencies obtainable.

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1. *Thermodynamics of Magnetic Substances*

The thermodynamic properties of magnetic substances have been investigated for many years in connection with anomalies of the specific heat near the Curie point[2] and, more recently, a very important use of these theories has been made in the magnetic method of obtaining extremely low temperatures,[3] down to 1/100 of a degree, absolute. A short account of the theory may be found in textbooks.[4] Similar results have also been obtained on dielectrics.[5] A complete statistical theory of magnetism appears also in texts.[6]

The thermodynamic properties of magnetic substances will be summarized and some useful formulas for determining the conditions for maximum efficiency in a thermomagnetic generator will be derived. To avoid unnecessary complications, we will assume that hysteresis effects do not exist, thus indicating complete thermodynamic reversibility.

When heat dQ is applied to a magnetic substance, varying its temperature T by dT and its induction B by dB, we have

$$dQ = c(T,B)dT + fdB, \tag{1}$$

where c = heat capacity per unit volume at constant induction B, f = coefficient of "heat of magnetization."

The magnetic work dW done by the magnetic system per unit volume is given[7] by

$$-dW = HdB, \tag{2}$$

where H is the total field.

According to the law of conservation of energy, the increase dU in the internal energy of the magnetic body is given by

$$dU = dQ + (-dW) = c(T,B)dT + (f+H)dB. \tag{3}$$

Equation (3) is an exact total differential; hence

$$\left.\begin{aligned} &\frac{\partial}{\partial B}\left(\frac{\partial U}{\partial T}\right) = \frac{\partial}{\partial T}\left(\frac{\partial U}{\partial B}\right) \\ \text{or}\quad & \\ &\left.\frac{\partial c}{\partial B}\right]_T = \left.\frac{\partial f}{\partial T}\right]_B + \left.\frac{\partial H}{\partial T}\right]_B . \end{aligned}\right\} \tag{4}$$

Equation (4) shows the dependence of c on B.

Since a reversible process has been assumed, the entropy change

$$dS = \frac{dQ}{T} = \frac{c}{T}dT + \frac{f}{T}dB \tag{5}$$

must also be an exact total differential. Hence,

$$\left.\begin{aligned} &\frac{\partial}{\partial B}\left(\frac{\partial S}{\partial T}\right) = \frac{\partial}{\partial T}\left(\frac{\partial S}{\partial B}\right) \\ \text{or}\quad & \\ &\frac{\partial}{\partial B}\left(\frac{c}{T}\right) = \frac{\partial}{\partial T}\left(\frac{f}{T}\right) . \end{aligned}\right\} \tag{6}$$

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[1] Thomas Edison, British Patent 16,709; 1887: N. Tesla, U.S. Patent 428,057; 1890: Chilowsky, U.S. Patent Applications 627,832 and 635,988.

[2] Weiss and Forrer, "Magneto-Caloric Effect," *Annales de Physique*, v. 15, pp. 153–213; January–February, 1926.

[3] Giauque and MacDougall, "The Production of Temperatures Below One Degree Absolute by Adiabatic Demagnetization of Gadolinium Sulphate," *Journal of the American Chemical Society*, v. 57, pp. 1175–1185; July, 1935.

[4] Zemansky, "Heat and Thermo Dynamics," 2nd Edition, McGraw-Hill Book Company, New York, New York, pp. 271–279.

[5] Brillouin, "Propagation des Ondes Electromagnetiques dans les Milieux Matériels," *Congress International d'Electricité*, Paris, 1932; v. II, pp. 739–786.

[6] Van Vleck, "Theory of Electric and Magnetic Susceptibility," 1st Edition, Oxford-Clarendon Press, 1932.

[7] J. A. Stratton, "Electromagnetic Theory," 1st Edition, McGraw-Hill Book Company, New York, New York, 1941; p. 124: Also, R. W. P. King, "Electromagnetic Engineering," v. 1, 1st Edition, McGraw-Hill Book Company, New York, New York, 1945; p. 187.

Hence,

$$\frac{1}{T}\left(\frac{\partial c}{\partial B}\right)_T=-\frac{f}{T^2}+\frac{1}{T}\left(\frac{\partial f}{\partial T}\right)_B$$

and

$$\left[\frac{\partial c(T,B)}{\partial B}\right]_T=-\frac{f}{T}+\left[\frac{\partial f}{\partial T}\right]_B. \tag{7}$$

From (4) and (7),

$$f=-T\left[\frac{\partial H}{\partial T}\right]_B \tag{8}$$

and

$$\left.\frac{\partial c(T,B)}{\partial B}\right]_T=-T\left[\frac{\partial^2 H}{\partial T^2}\right]_B; \tag{9}$$

hence, by integration of (9),

$$c(T,B)=c(T,O)-T\int\left[\frac{\partial^2 H}{\partial T^2}\right]dB. \tag{10}$$

According to (3) this.... $=\left[\dfrac{\partial U}{\partial T}\right]_B$.

From (1) and (8), the heat given to the system becomes

$$dQ=TdS=c(T,B)dT-T\left[\frac{\partial H}{\partial T}\right]dB. \tag{11}$$

We also have, from (3) and (8),

$$\frac{\partial U}{\partial B}=H-T\left[\frac{\partial H}{\partial T}\right]_B. \tag{12}$$

The factors T and B were taken as the independent variables in the foregoing discussion. We may change to variables T and H, instead, by taking $B(T,H)$ or

$$dB=\left[\frac{\partial B}{\partial T}\right]_H dT+\left[\frac{\partial B}{\partial H}\right]_T dH. \tag{13}$$

Then, corresponding to (1), we have the expression

$$dQ=c(T,H)dT+gdH, \tag{14}$$

where $c(T,H)$ denotes heat capacity at a constant field H.

Starting from (2) and (13), we have, for work done by the system,

$$-dW=HdB=H\left(\frac{\partial B}{\partial T}\right)_H dT+H\left(\frac{\partial B}{\partial H}\right)_T dH. \tag{15}$$

By an analogous reasoning, writing conditions that make dU and dS exact total differentials, we obtain

$$g=T\left[\frac{\partial B}{\partial T}\right]_H. \tag{16}$$

$$\left.\begin{aligned}TdS=dQ&=c(T,H)dT+T\left(\frac{\partial B}{\partial T}\right)_H dH,\\ \left[\frac{\partial c(T,H)}{\partial H}\right]_T&=T\left[\frac{\partial^2 B}{\partial T^2}\right]_H,\\ dU&=cdT+(g+B)dH,\end{aligned}\right\} \tag{17}$$

$$c(T,H)=c(T,O)+T\int_0^H\left(\frac{\partial^2 B}{\partial T^2}\right)_H dH, \tag{18}$$

and

$$\left.\frac{\partial U}{\partial H}\right]_T=T\left[\frac{\partial B}{\partial T}\right]_H+H\left[\frac{\partial B}{\partial H}\right]_T. \tag{19}$$

It should be noted that these relations do not imply any special assumption about the relations between B, H, and T.

2. *Simplifications for Substances Having Constant Permeability*

2.1 $B=\mu(T)H$

If we consider an average value of permeability μ dependent on temperature only,

$$H(T,B)=\frac{1}{\mu(T)}B, \tag{20}$$

(3) and (5), with the help of (8), yield

$$U(T,B)=U(T,O)+\tfrac{1}{2}B^2\left(\frac{1}{\mu}-T\frac{\partial\frac{1}{\mu}}{\partial T}\right) \tag{21}$$

and

$$S(T,B)=S(T,O)-\tfrac{1}{2}B^2\frac{\partial\frac{1}{\mu}}{\partial T}. \tag{22}$$

Equation (10) then gives

$$c(T,B)=c(T,O)-\tfrac{1}{2}B^2T\frac{\partial^2\frac{1}{\mu}}{\partial T^2}, \tag{23}$$

where

$$c(T,O)=\frac{\partial U(T,O)}{\partial T}=T\frac{\partial S(T,O)}{\partial T} \tag{24}$$

is the heat capacity for zero induction. Equation (11) now becomes

$$dQ=c(T,B)dT-T\left[\frac{\partial\frac{1}{\mu}}{\partial T}\right]_B BdB. \tag{25}$$

This equation gives, for $dT=0$, the amount of heat to be given to the system during a change of magnetization dB.

Turning now to variables T,H as in (13) to (18), by integration of dU,

$$U(T,H)=U(T,O)+\tfrac{1}{2}H^2\left(\mu+T\frac{\partial\mu}{\partial T}\right) \quad (26)$$

and

$$c(T,H)=c(T,O)+\tfrac{1}{2}H^2T\frac{\partial^2\mu}{\partial T^2}, \quad (27)$$

with

$$dQ=c(T,H)dT+T\frac{\partial\mu}{\partial T}HdH. \quad (28)$$

For paramagnetic substances that follow the Curie-Weiss law,

$$\mu=\mu_0+\frac{C}{T-\Theta}, \quad (29)$$

where $T>\Theta$ (Curie temperature), and μ_0, the permeability of a vacuum, may be neglected. From (29),

$$\frac{\partial}{\partial T}\frac{1}{\mu}=\frac{1}{C}$$

and

$$\frac{\partial^2}{\partial T^2}\frac{1}{\mu}=0.$$

Equations (25) and (28) become

$$\begin{aligned}dQ&=c(T,B)dT-\frac{T}{C}BdB\\&=c(T,H)dT-\frac{CT}{(T-\Theta)^2}HdH.\end{aligned} \quad (30)$$

The heat-capacity equations given by (23) and (27) become

$$c(T,B)=c(T,O)$$

and

$$c(T,H)=c(T,O)+H^2T\frac{C}{(T-\Theta)^3}. \quad (31)$$

As is known, the usual heat capacity $c(T,O)$ of solids drops to negligible values at very low temperatures of the order of T^3. During an adiabatic demagnetization, it follows from (30) and (31),

$$\begin{aligned}c(T,H)dT&=H^2T\frac{C}{(T-\Theta)^3}dT\\&=\frac{CT}{(T-\Theta)^2}HdH.\end{aligned}$$

Hence,

$$\frac{dT}{T-\Theta}=\frac{dH}{H} \quad (32)$$

or

$$T-\Theta=AH, \quad (33)$$

where $A=$ a constant. By reducing H to zero, the temperature drops to the Curie point Θ. Then, from (29) and (33),

$$B=H=\frac{C}{T-\Theta}\times\frac{T-\Theta}{A}=\frac{C}{A},$$

meaning that B remains constant during this adiabatic process. Some salts, like gadolinium sulfate, have a Curie temperature very near absolute zero. Very low temperatures may be reached by this adiabatic demagnetization process.[3]

2.2 $B(T,H)=B(T,O)+\mu_\Delta H$

If we assume the soft ferromagnetic component of the generator to be nearly saturated throughout a heat cycle (only a few oersteds are required), we may write

$$B(T,H)=B(T,O)+\mu_\Delta H. \quad (34)$$

For this case, expressions for internal energy $U(T,H)$, heat capacity $c(T,H)$, and entropy $S(T,H)$ become as follows from (17), (18), (19), and (34):

$$\begin{aligned}U(T,H)=U(T,O)&+HT\frac{\partial B(T,O)}{\partial T}\\&+\tfrac{1}{2}H^2\left(\mu_\Delta+T\frac{\partial\mu_\Delta}{\partial T}\right),\end{aligned} \quad (35)$$

$$\begin{aligned}c(T,H)=c(T,O)&+HT\frac{\partial^2B}{\partial T^2}(T,O)\\&+\tfrac{1}{2}H^2T\frac{\partial^2\mu_\Delta}{\partial T^2},\end{aligned} \quad (36)$$

and

$$S(T,H)=S(T,O)+H\frac{\partial B(T,O)}{\partial T}+\tfrac{1}{2}H^2\frac{\partial\mu_\Delta}{\partial T}, \quad (37)$$

where

$$c(T,O)=\frac{\partial U(T,O)}{\partial T}=T\frac{\partial S(T,O)}{\partial T}$$

is the heat capacity for zero induction. Equation (17) gives

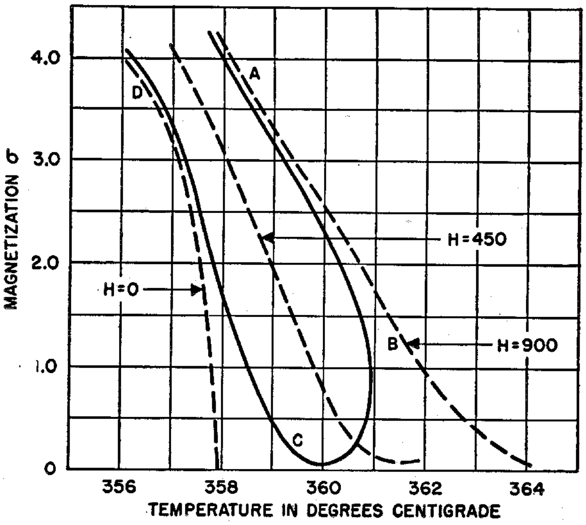

Fig. 1—Magnetization-versus-temperature curve for heat cycle *A–B–C–D* shown by solid line. Values of *H* are in oersteds. $(\overline{\partial\sigma}/\partial T)\simeq 1.2$ per heat cycle. Based on experimental data for nickel by Weiss and Forrer.

$$
\begin{aligned}
dQ&=TdS\\
&=c(T,H)dT+T\left[\frac{\partial B(T,O)}{\partial T}+H\frac{\partial \mu_\Delta}{\partial T}\right]dH \qquad (38)
\end{aligned}
$$

as the amount of heat required by the system to maintain its temperature constant during magnetization.

3. *Ideal Thermomagnetic Generator; Carnot and T,H Cycles*

In a thermomagnetic generator, the sample is acted on by a permanent magnet and a coil through which a current i flows, contributing to a magnetic field H, while the temperature is varied. Hence, it seems convenient to use T and H as independent variables; we shall assume here $B(T,H)=\mu(T)H$. Maximum efficiency should be attained with a Carnot cycle consisting of two adiabatic branches connecting two branches at T_1 and T_2.

First, consider an adiabatic transformation with $dQ=0$. According to (27) and (28),

$$\left[c(T,O)+\tfrac{1}{2}H^2T\frac{\partial^2\mu}{\partial T^2}\right]dT+T\frac{\partial\mu}{\partial T}HdH=0. \qquad (39)$$

A rise dT in temperature is accompanied by an increase dH in magnetic field.

We intend to work with a ferromagnetic substance in which permeability decreases when T is increased and becomes μ_0 at the Curie temperature Θ. This suggests a law of the form

$$\mu=\mu_0+A(\Theta-T)^n, \qquad (40)$$

where $A>0$, $T\leqslant\Theta$, and the exponent n seems to be of the order of 1 or 2.

Experimental data obtained by Weiss and Forrer on nickel show, in Fig. 1, a dependence of σ on H in addition to T. Since this leads to great complications, we shall take average values of A and Θ that correspond to an average field intensity and, using an exponent $n=1$ (Fig. 2),

$$\left.\begin{aligned}
&\mu=\mu_0+A(\Theta-T),\\
&\frac{\partial\mu}{\partial T}=-A,\\
&\frac{\partial^2\mu}{\partial T^2}=0,
\end{aligned}\right\} \qquad (41)$$

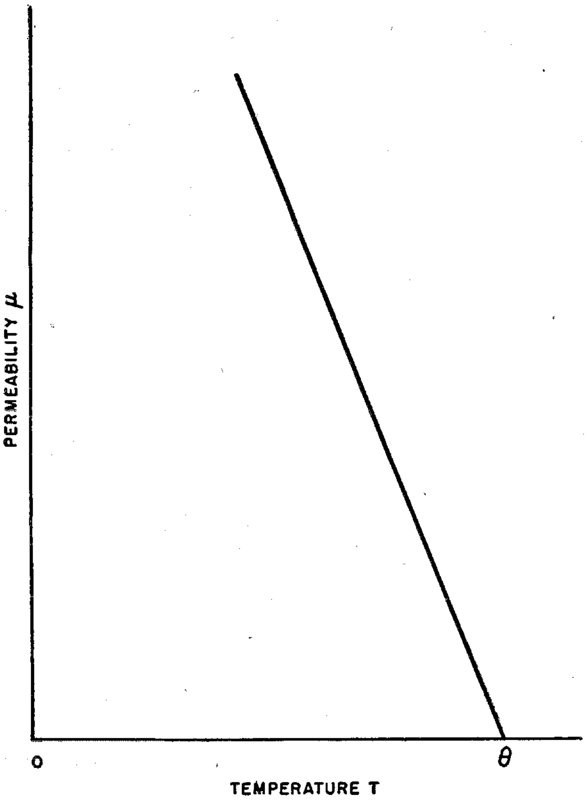

Fig. 2—Permeability μ plotted against temperature for a thermomagnetic generator. Θ is the Curie temperature. The slope of the curve $=-A$.

where $T \leqslant \Theta$. Then, from (27) and (39),

$$\left.\begin{array}{l} c(T,H) = c(T,O), \\ c(T,O)dT = ATH\,dH, \end{array}\right\} \tag{42}$$

for an adiabatic transformation. The heat capacity $c(T,O)$ can be treated as a constant c_0 and (42) is integrated:

$$c_0 \log \frac{T}{T_0} = \tfrac{1}{2}AH^2, \tag{43}$$

where T_0 is the temperature for which $H=0$.

$$T = T_0\, e \exp\left[\frac{A}{2c_0}H^2\right], \tag{44}$$

where $T \leqslant \Theta$. The Carnot cycle is approximately plotted in Fig. 3. Thermodynamic efficiency $\mathcal{E}$ is defined as the ratio

$$\mathcal{E} = \frac{W}{Q_1}, \tag{45}$$

which is work output W per cycle for a heat input Q_1, not counting the amount of heat $-Q_2$ returned to the source at a lower temperature T. On the other hand,

$$W = Q_1 - |Q_2| = Q_1 + Q_2 \tag{46}$$

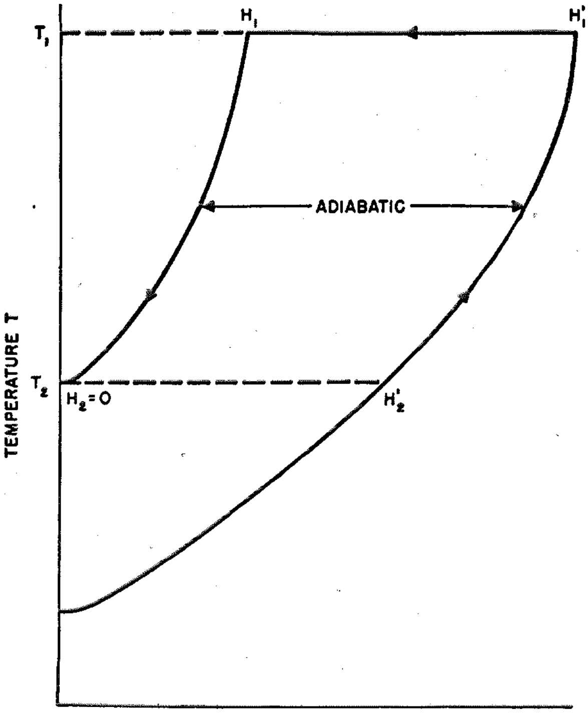

Fig. 3—Carnot cycle for a thermomagnetic generator.

in consideration of energy conservation; hence,

$$\mathcal{E} = 1 - \frac{Q_2}{Q_1}. \tag{47}$$

In this example, we can easily compute Q_1 and Q_2 from (28). During an isothermal transformation (at either T_1 or T_2), the amount of heat input is

$$Q = \int_H^{H'} dQ = T\frac{\partial\mu}{\partial T}\int_H^{H'} H\,dH$$
$$= \tfrac{1}{2}T\frac{\partial\mu}{\partial T}(H'^2 - H^2), \tag{48}$$

or, according to (41),

$$Q = -\tfrac{1}{2}AT(H'^2 - H^2), \tag{49}$$

where $T \leqslant \Theta$. From the cycle of Fig. 3 and using (43) for the adiabatic branches, we have

$$c_0 \log \frac{T_1}{T_2} = \tfrac{1}{2}AH_1^2 = \tfrac{1}{2}A(H_1'^2 - H_2'^2),$$

and thus $H_1^2 - H_1'^2 = -H_2'^2$. For the highest temperature T_1, (49) yields

$$Q_1 = \tfrac{1}{2}AT_1(H_1^2 - H_1'^2) \tag{50}$$

for heat input,

$$Q_2 = -\tfrac{1}{2}AT_2H_2'^2$$

for heat output at the lowest temperature T_2, and the efficiency by (50) is

$$\mathcal{E} = 1 - \frac{|Q_2|}{Q_1} = 1 - \frac{T_2}{T_1} = \frac{T_1 - T_2}{T_1}. \tag{51}$$

This is the expected maximum efficiency for a Carnot cycle.

Another cycle to be considered is the T,H cycle shown in Fig. 4. The heat given to the system along the isothermal branches at T_1 or T_2 can be computed from (49). In addition, we have to consider the heating process AB at a constant field H' and the cooling CO under zero field. The heat input is

$$Q_1 = Q_{AB} + Q_{BC} = c(T,H)(T_1 - T_2) + \frac{AT_1}{2}H'^2. \tag{52}$$

The heat output is

$$|Q_2| = |Q_{CO}| + |Q_{OA}|$$
$$= c(T,O)(T_1 - T_2) + \frac{AT_2}{2}H'^2, \tag{53}$$

and according to (42), $c(T,H)=c(T,O)=c_0$. The efficiency becomes

$$\left.\begin{aligned}\mathcal{E}_{TH}&=1-\frac{|Q_2|}{Q_1}\\&=1-\frac{\frac{1}{2}AT_2H'^2+c_0(T_1-T_2)}{\frac{1}{2}AT_1H'^2+c_0(T_1-T_2)}\end{aligned}\right\} \qquad (54)$$

and is obviously smaller than the efficiency for the Carnot cycle. If the heat capacity c_0 be large enough, the efficiency may become very small, as seen from simplification of (54).

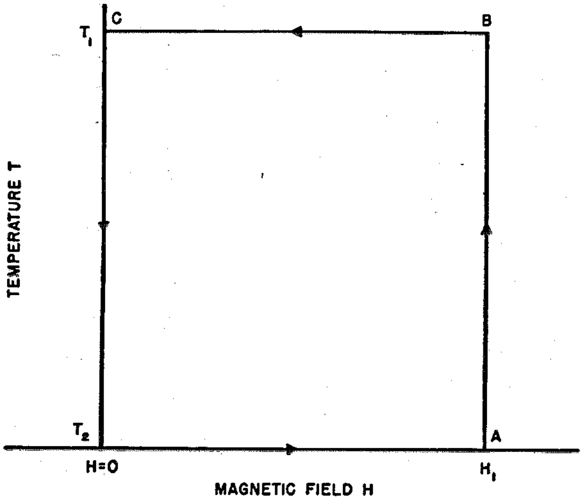

Fig. 4—*T,H* cycle for a thermomagnetic generator.

$$\mathcal{E}_{TH}=\frac{T_1-T_2}{T_1+\dfrac{2c_0}{AH'^2}(T_1-T_2)}. \qquad (55)$$

4. *Actual Thermomagnetic Generator and Its Cycle for Case of B(T,H) = μ(T)H*

4.1 OPTIMUM CURRENT AND POWER

Assume a structure like that represented in Fig. 5, with a magnetically soft[8] component of variable temperature T in a field H with a large average component H_0 due to a permanent magnet, plus the field induced by the current i in the coils:

$$H=H_0+n_0i \qquad (56)$$

in meter-kilogram-second rational units, where n_0 = number of turns of the coil per unit length, $n=n_0l$ = total number of turns for a coil of length l. The temperature is supposed to vary sinusoidally,

$$T=T_a+b\sin\omega t. \qquad (57)$$

Highest temperature $=T_1=T_a+b\leqslant\Theta$.
Lowest temperature $=T_2=T_a-b$.

The permeability consequently follows a similar law:

$$\begin{aligned}\mu&=\mu_a-Ab\sin\omega t,\\ \mu_a&=\mu(T_a),\end{aligned} \qquad (58)$$

are average values assuming for $\mu(T)$ the same temperature dependence as in (41) and taking for A the average value of $-(\partial\mu/\partial T)$ over a whole cycle. The magnetic induction is

$$B=\mu H=(\mu_a-Ab\sin\omega t)(H_0+n_0i). \qquad (59)$$

Let us assume the coil to be connected to a resistance R and write the electrical equation of the circuit. Calling $\phi=SB$ = the flux through a turn of surface S, we have

$$n\frac{d\phi}{dt}=nS\frac{dB}{dt}=-Ri, \qquad (60)$$

or, according to (59),

$$\begin{aligned}\mu_ann_0S\frac{di}{dt}-nn_0SAb\frac{d}{dt}(i\sin\omega t)+Ri&\\=nSAbH_0\cos\omega t.& \qquad (61)\end{aligned}$$

In the first term of this equation, we recognize the self-induction of the circuit,

$$L=\mu_ann_0S. \qquad (62)$$

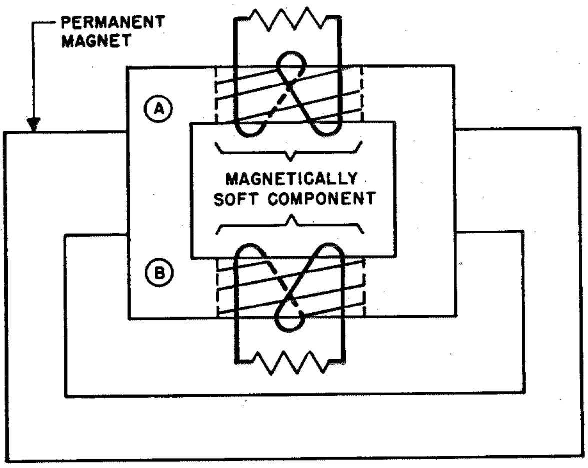

Fig. 5—Magnetic circuit of a thermomagnetic generator.

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[8] *Magnetically soft* refers to materials with high μ's, in contrast to *magnetically hard* materials used for permanent magnets.

Assuming that the magnetic field due to the current i remains small[9] compared with H_0,

$$n_0 i \ll H_0, \tag{63}$$

we may ignore the second term as being small compared with the right-hand term, and obtain

$$L\frac{di}{dt}+Ri=\beta\omega\cos\omega t, \tag{64}$$

where $\beta=nSAbH_0$. We are not interested in the transients at the start of the operation, and look only for the final steady-state oscillations.

$$i=Ie^{j\omega t}, \tag{65}$$

where $I=\dfrac{\beta\omega}{j\omega L+R}$, which becomes maximum for $R=\omega L\ldots.|I_{\max}|=\beta/2^{\frac{1}{2}}L$. The power output is

$$P=\tfrac{1}{2}R|I|^2=\tfrac{1}{2}\left(\frac{\beta^2\omega^2R}{L^2\omega^2+R^2}\right), \tag{66}$$

and becomes maximum when

$$\left.\begin{aligned} R&=L\omega,\\ P_{\max}&=\frac{\beta^2\omega}{4L}\\ &=\frac{\omega}{4}\frac{nS}{\mu_a n_0}(AbH_0)^2\\ &=\frac{\omega V}{4\mu_a}(AbH_0)^2, \end{aligned}\right\} \tag{67}$$

according to (62) and (64), where $n=n_0 l$ and $V=Sl$ is the volume of the sample.

If the highest temperature is the Curie temperature Θ, for which μ is practically zero, we must state

$$\mu(T_1)=\mu(\Theta)=\mu_a-Ab=0 \tag{68}$$

in (58), and (67) then reduces to

$$\left.\begin{aligned} P_{\max}&=\frac{(\omega)}{4}V\mu_a H_0^2,\\ \text{and}\qquad T_1&=T_a+b=\Theta, \end{aligned}\right\} \tag{69}$$

where $\mu_a=\mu(T_a)=Ab=\tfrac{1}{2}\mu(T_2)=\tfrac{1}{2}\mu_{\max}$. The work done per cycle and per unit volume is

$$\frac{W}{Vf}=\frac{\pi}{2}\mu_a H_0^2. \tag{70}$$

4.2 Efficiency of Generator

We must now investigate the heat input of the system at different phases of a complete cycle to compute the efficiency. Let us assume the optimum resistance (67) in the electric circuit that will result in a current (65);

$$i=\frac{1}{L(1+j)}e^{j\omega t}=\frac{\beta}{2^{\frac{1}{2}}L}e\exp\left[j\left(\omega t-\frac{\pi}{4}\right)\right],$$

or

$$\left.\begin{aligned} i&=\frac{\beta}{2^{\frac{1}{2}}L}\cos\left(\omega t-\frac{\pi}{4}\right)\\ &=\frac{\beta}{2^{\frac{1}{2}}L}\sin\left(\omega t+\frac{\pi}{4}\right). \end{aligned}\right\} \tag{71}$$

Hence,

$$H=H_0+\frac{\beta n_0}{2^{\frac{1}{2}}L}\sin\left(\omega t+\frac{\pi}{4}\right)$$

according to (56). Referring to (62) and (64), we have

$$\begin{aligned} \frac{\beta n_0}{2^{\frac{1}{2}}L}&=\frac{nn_0SAbH_0}{2^{\frac{1}{2}}\mu_a nn_0 S}\\ &=\frac{Ab}{2^{\frac{1}{2}}\mu_a}H_0\\ &=\frac{1}{2^{\frac{1}{2}}}H_0, \end{aligned}$$

when the highest temperature is the Curie point

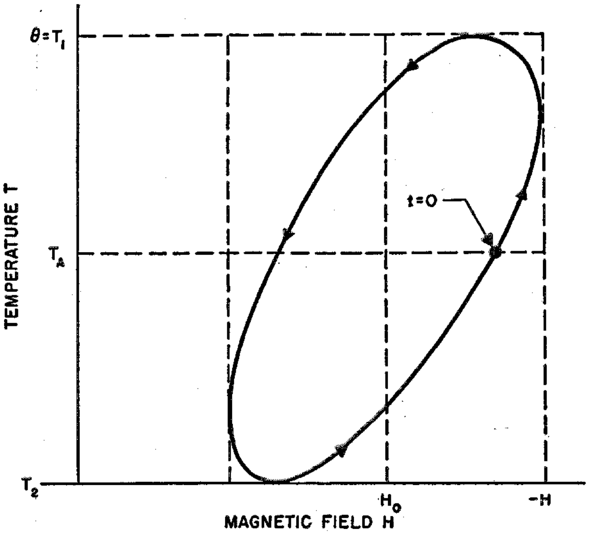

Fig. 6—Magnetizing field plotted against temperature.

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[9] Retaining $d/dt(i\sin\omega t)$ in the equation does not materially change the practical results of the discussion but introduces higher harmonics in the current.

Θ and (68) is satisfied. Then,

$$H=H_0\left[1+\frac{1}{2^{\frac{1}{2}}}\sin\left(\omega t+\frac{\pi}{4}\right)\right]. \qquad (72)$$

Comparing (72) and (57), we obtain the H,T cycle plotted in Fig. 6. If we compare (72) with (59) and (63), we see that the variable term, representing the field due to current i, is not very small ($H_0/2^{\frac{1}{2}}$ compared with H_0). A more detailed calculation shows, however, that this only introduces harmonics in the current without much change in the main contribution. For a small transformation dT, dH (equation (28)),

$$dQ=c(T,H)dT+T\frac{\partial\mu}{\partial T}HdH,$$

gives the heat input per unit volume. Or

$$dQ=c(T,O)dT-ATHdH, \qquad (73)$$

when we use the magnetization law of (41). We can treat the heat capacity $c(T,O)$ as a constant c_0, and use (57) and (72);

$$\left.\begin{aligned}\frac{dQ}{dt}&=c_0\frac{dT}{dt}-\tfrac{1}{2}AT\frac{d(H^2)}{dt}\\ &=c_0b\omega\cos\omega t\\ &\qquad-\tfrac{1}{2}AT_aH_0^2\frac{d}{dt}\left(1+\frac{1}{2^{\frac{1}{2}}}\sin\phi\right)^2\\ &\qquad-\tfrac{1}{2}Ab\sin\omega tH_0^2\frac{d}{dt}\left(1+\frac{1}{2^{\frac{1}{2}}}\sin\phi\right)^2,\end{aligned}\right\} \qquad (74)$$

where we have $\phi=\omega t+(\pi/4)$; now,

$$\frac{d}{dt}\left(1+\frac{1}{2^{\frac{1}{2}}}\sin\phi\right)^2=\omega 2^{\frac{1}{2}}\cos\phi\left(1+\frac{1}{2^{\frac{1}{2}}}\sin\phi\right)$$
$$=\omega 2^{\frac{1}{2}}\cos\phi+\frac{\omega}{2}\sin 2\phi.$$

Then,

$$\begin{aligned}\frac{1}{\omega}\frac{dQ}{dt}&=c_0b\cos\omega t-\frac{A}{2^{\frac{1}{2}}}T_aH_0^2\cos\phi\\ &\quad-\frac{A}{4}T_aH_0^2\sin 2\phi-\frac{A}{2^{\frac{1}{2}}}bH_0^2\sin\omega t\cos\phi\\ &\quad-\frac{A}{4}bH_0^2\sin\omega t\sin 2\phi.\end{aligned} \qquad (75)$$

This equation may be put in the form

$$\begin{aligned}\frac{1}{\omega}\frac{dQ}{dt}&=\frac{A}{4}bH_0^2\\ &\quad+c_0b\cos\omega t-\frac{A}{2^{\frac{1}{2}}}T_aH_0^2\cos\phi+\frac{A}{8}bH_0^2\sin\omega t\\ &\quad-\frac{A}{4}T_aH_0^2\sin 2\phi-\frac{AbH_0^2}{2\times 2^{\frac{1}{2}}}\sin\left(2\omega t+\frac{\pi}{4}\right)\\ &\quad-\frac{A}{8}bH_0^2\sin 3\omega t.\end{aligned} \qquad (76)$$

The net input for a complete cycle is

$$q=\int_{t=0}^{t=2\pi/\omega}\frac{dQ}{dt}dt=A\frac{\pi}{2}bH_0^2, \qquad (77)$$

since only the first constant term contributes to the integral taken over a whole period. This corresponds exactly to (70), relating to the work done by the system per cycle for energy conservation.

As in the examples discussed in the preceding section, we must distinguish between the heat input Q_1 during one half of the cycle, and the heat output $Q_2=-|Q_2|$ during the other half. The net input q is

$$q=Q_1+Q_2=Q_1-|Q_2|. \qquad (78)$$

To compute the heat input, we integrate from $\omega t=\theta$ to $\omega t=\theta+\pi$, and we choose θ such as to make Q_1 maximum and positive.[10] The integration of terms in 2ϕ or $2\omega t$ obviously gives zero. Replacing ϕ by $\omega t+\pi/4$, and expanding and regrouping terms, we obtain

$$\begin{aligned}Q_1=\tfrac{1}{2}q+b\int_\theta^{\theta+\pi}&\left[\left(c_0-\frac{T_a}{2b}AH_0^2\right)\cos\omega t\right.\\ &\left.+\left(\frac{1}{8}+\frac{T_a}{2b}\right)AH_0^2\sin\omega t-\tfrac{1}{8}AH_0^2\sin 3\omega t\right]dt\end{aligned} \qquad (79)$$

or

$$\begin{aligned}Q_1=\tfrac{1}{2}q+b&\left[\left(2c_0-\frac{T_a}{b}AH_0^2\right)\sin\theta\right.\\ &\left.-\left(\frac{1}{4}+\frac{T_a}{b}\right)AH_0^2\cos\theta+\tfrac{1}{12}AH_0^2\cos 3\theta\right].\end{aligned}$$

---

[10] This means that we choose a half cycle during which heat is given to the system, while the other half cycle, from $\theta+\pi$ to $\theta+2\pi$, corresponds to heat released by the system.

The last term is small and may be dropped. Then,

$$Q_1 \simeq \tfrac{1}{2}q + b\left[\left(2c_0 - \frac{T_a}{b}AH_0^2\right)^2 + \left(\frac{1}{4} + \frac{T_a}{b}\right)^2 A^2H_0^4\right]^{\frac{1}{2}}. \qquad (80)$$

The thermodynamic efficiency is given by

$$\begin{aligned}\mathcal{E} &= \frac{q}{Q_1}\\ &= \frac{q}{\tfrac{1}{2}q + \left[(2c_0b - T_aAH_0^2)^2 + \left(\frac{b}{4} + T_a\right)^2 A^2H_0^4\right]^{\frac{1}{2}}}\\ &= \frac{\pi/2}{\frac{\pi}{4} + [(D-X)^2 + (\frac{1}{4}+X)^2]^{\frac{1}{2}}}, \qquad (81)\end{aligned}$$

using (77) for q and putting $D = 2c_0/AH_0^2$ and $X = T_a/b$. We must discuss the variation in efficiency when the temperature variation $\Delta T = 2b$ is continuously changed from 0 to its maximum (and unattainable) value $2T_a$, when $b = T_a$. This corresponds to a variation of X from infinity to 1.

4.2.1 *Case I*

If $b \ll T_a/D$ and $X \gg D$,

$$\mathcal{E}_{\mathrm{I}} = \frac{\frac{1}{2}\pi}{\frac{1}{4}\pi + 2^{\frac{1}{2}}X} = \frac{2\pi b}{\pi b + 4\times 2^{\frac{1}{2}}T_a} \simeq \frac{\pi\Delta T}{4\times 2^{\frac{1}{2}}T_1}. \qquad (82)$$

Then, efficiency amounts to $\pi/4\times 2^{\frac{1}{2}} = 0.555$ times the corresponding Carnot efficiency, $\mathcal{E}_c = \Delta T/T_1$.

4.2.2 *Case II*

If $b = T_a/D$ and $X = D$,

$$\left.\begin{aligned}\mathcal{E}_{\mathrm{II}} &= \frac{\frac{\pi}{2}}{\frac{\pi}{4} + X + \frac{1}{4}}\\ &= \frac{2\pi}{\pi + 1 + 4D}\end{aligned}\right\} \qquad (83)$$

maximum. When b is increased further, the radical goes through a minimum that corresponds to maximum efficiency:

$$D - X = \tfrac{1}{4} + X,$$

$$X = \tfrac{1}{2}D - \tfrac{1}{8}, \quad \text{or} \quad \Delta T \simeq \frac{4T_a}{D}, \qquad (84)$$

$$(D-X)^2 + (\tfrac{1}{4}+X)^2 = \tfrac{1}{2}(D+\tfrac{1}{4})^2,$$

and

$$\mathcal{E}_{\max} = \frac{\frac{\pi}{2}}{\frac{\pi}{4} + \frac{1}{2^{\frac{1}{2}}}(D+\frac{1}{4})} = \frac{2\pi}{\pi + \frac{1}{2^{\frac{1}{2}}} + 2\times 2^{\frac{1}{2}}D} \simeq \frac{\pi}{2^{\frac{1}{2}}D} \qquad (85)$$

when D is large.

4.2.3 *Case III*

Larger values of b, if they can be obtained, yield a finite efficiency; if $b \gg T_a/D$, $X \ll D$,

$$\mathcal{E}_{\mathrm{III}} = \frac{\pi/2}{\pi/4 + D} = \frac{2\pi}{\pi + 4D} \simeq \frac{\pi}{2D}$$

for large D values, and the variation of efficiency with respect to b is represented by the curve of Fig. 7. The question of whether the maximum can be reached depends on its position, for we always have the *limitation*

$$\frac{1}{X} = \frac{b}{T_a} \leqslant 1. \qquad (86)$$

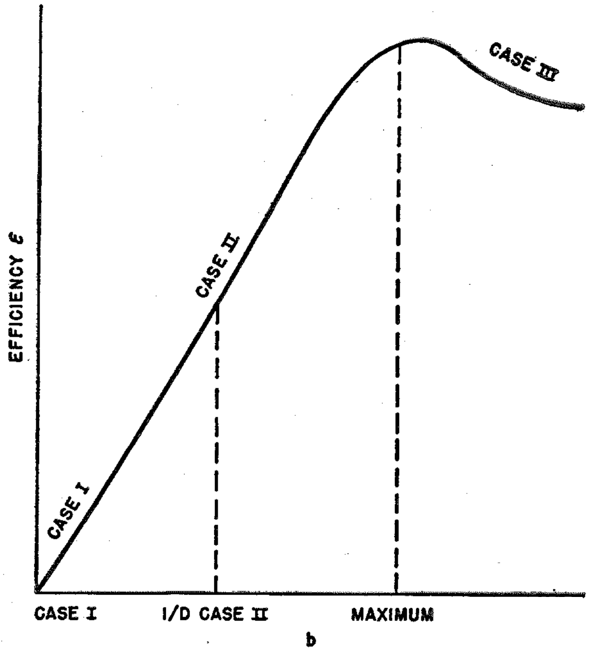


Fig. 7—Efficiency $\mathcal{E}$ plotted against b; $1/X = b/T_a$.

Much will depend on the actual value of the coefficient

$$D = \frac{2c_0}{AH_0^2},$$

where c_0 = heat capacity per unit volume under zero magnetic field,

$-A = \partial\mu/\partial T$ = average rate of change of $\mu(T)$, and

H_0 = average magnetic field from permanent magnet.

This coefficient D is a dimensionless quantity that can be written

$$\left.\begin{aligned} D &= \frac{2c_0\Delta T}{\Delta\mu H_0^2} = \frac{\Delta Q}{\Delta W_m}, \\ \Delta Q &= c_0 \Delta T, \\ \Delta W_m &= \tfrac{1}{2}\Delta\mu H_0^2, \end{aligned}\right\} \tag{88}$$

in meter-kilogram-second units, where

ΔQ = caloric energy needed to increase the temperature of a certain volume of matter by ΔT,

ΔW_m = corresponding increase of magnetic energy.

5. *Actual Thermomagnetic Generator and Its Cycle for $B(T,H) = B(T,O) + \mu_\Delta H$*

5.1 OPTIMUM CURRENT AND POWER

We may derive expressions for optimum power and input energy in a manner similar to the preceding case. In (34), μ_Δ is taken to be nearly independent of temperature; hence,

$$\frac{\partial B(T,H)}{\partial T} \simeq \frac{\partial B(T,O)}{\partial T}.$$

In centimeter-gram-second electromagnetic units,

$$B(T,O) = 4\pi\rho\sigma(T). \tag{89}$$

We may write

$$\sigma_T \simeq k(\Theta - T), \tag{90}$$

where k = a constant. Then, $B(T,O) = 4\pi\rho k(\Theta - T)$ in centimeter-gram-second electromagnetic units, or

$$B(T,O) = K(\Theta - T) \tag{91}$$

in meter-kilogram-second units if $K = 4\pi\rho \times k10^{-4}$. In the meter-kilogram-second system,

$$\mu_\Delta = h\mu_0, \tag{92}$$

where h has a magnitude of a few units and $\mu_0 = 1.26 \times 10^{-6}$. Hence,

$$B(T,H) = B(T,O) + \mu_\Delta H = K(\Theta - T) + h\mu_0 H \tag{93}$$

as shown in Fig. 8. Using these values in (36) and (38), we obtain $c(T,H) = c(T,O)$, and

$$dQ = c(T,O)dT - KTdH. \tag{94}$$

The equation of the generator will be

$$E = nS\frac{dB}{dt}(T,H), \tag{95}$$

where

$$B(T,H) = B(T,O) + \mu_\Delta(H_0 + n_0 i) \tag{96}$$

and

$$E = nS\frac{dB(T,O)}{dT}\frac{dT}{dt} + Sn\mu_\Delta n_0\frac{di}{dt}. \tag{97}$$

Assuming $T = T_a + b \sin \omega t$, we have

$$E = -nSKb\omega \cos \omega t + L\frac{di}{dt} = -Ri, \tag{98}$$

and, finally,

$$L\frac{di}{dt} + Ri = \beta'\omega \cos \omega t, \tag{99}$$

$\beta' = nSKb$.

This equation is identical in form to (64); however, there is no restriction such as (63), i.e., the only requirement is that

$$n_0 i \leqslant H_0. \tag{100}$$

The solution of (99) is given by

$$i = Ie \exp[j\omega t],$$

where $I = \dfrac{\beta'\omega}{j\omega L + R}$.

Again, for optimum conditions, $\omega L = R$, and we obtain (see (65) and (67))

$$\left.\begin{aligned} I_{\text{opt}} &= \frac{\beta'}{2^{\frac{1}{2}}L}, \\ i &= I_{\text{opt}} \sin\left(\omega t + \frac{\pi}{4}\right), \end{aligned}\right\} \tag{101}$$

and

$$P_{\text{opt}} = \frac{\omega V}{4\mu_\Delta}(Kb)^2. \tag{102}$$

Also, in view of (99) and (100),

$$n_0 I_{\text{opt}} = \frac{Kb}{2^{\frac{1}{2}}\mu_\Delta} \leqslant H_0.$$

Now, since it is desirable to increase $n_0 I$ as much as possible, we take $n_0 I_{\text{opt}} = H_0$, and the total H becomes

$$H = H_0\left[1 + \sin\left(\omega t + \frac{\pi}{4}\right)\right]. \tag{103}$$

It is to be noted that the maximum current here will be $2^{\frac{1}{2}}$ times greater than for the preceding case of (72).

Setting $\phi = \omega t + \frac{\pi}{4}$ in (103),

$$H = H_0(1 + \sin\phi),$$

and

$$dH = \omega H_0 \cos\phi\, dt.$$

5.2 Efficiency

The heat input for unit volume is given by (94) and, with the aid of (57),

$$\left.\begin{aligned}\frac{dQ}{dt} &= c(T,O)\frac{dT}{dt} - KT\frac{dH}{dt}\\ &= c_0 b\omega\cos\omega t\\ &\qquad - K(T_a + b\sin\omega t)\omega H_0\cos\phi,\end{aligned}\right\} \tag{104}$$

or

$$\frac{1}{\omega}\frac{dQ}{dt} = c_0 b\cos\omega t - KT_a H_0\cos\phi$$
$$- KbH_0\sin\omega t\cos\phi,$$

or

$$\frac{1}{\omega}\frac{dQ}{dt} = c_0 b\cos\omega t - KT_a H_0\cos\phi$$
$$+ KbH_0\left[\frac{1}{2\times 2^{\frac{1}{2}}} - \tfrac{1}{2}\sin\left(2\omega t + \frac{\pi}{4}\right)\right]. \tag{105}$$

The net input for a cycle is

$$q = \int_{t=0}^{2\pi/\omega}\left(\frac{dQ}{dt}\right)dt = \frac{\pi}{2^{\frac{1}{2}}}KbH_0. \tag{106}$$

To compute the heat input, we again integrate over a half-period θ, $\theta + \pi$, where θ is chosen so as to make Q_1 maximum and positive; this means that we choose the half-cycle when heat flows in and exclude the other half-cycle in which heat is released.

By carrying out this integration,

$$Q_1 = \tfrac{1}{2}q + c_0 b[\sin(\theta+\pi) - \sin\theta]$$
$$- KT_a H_0\left[\sin\left(\theta + \pi - \frac{\pi}{4}\right) - \sin\left(\theta + \frac{\pi}{4}\right)\right]$$
$$= \tfrac{1}{2}q - 2c_0 b\sin\theta + 2KT_a H_0\sin\left(\theta + \frac{\pi}{4}\right),$$
$$= \tfrac{1}{2}q + 2[(c_0 b)^2 + (KT_a H_0)^2$$
$$- 2^{\frac{1}{2}} c_0 b K T_a H]^{\frac{1}{2}}. \tag{107}$$

Now the thermodynamic efficiency of the generator becomes

$$\mathcal{E} = \frac{q}{Q_1}$$
$$= \frac{q}{\tfrac{1}{2}q + 2[(c_0 b)^2 + (KT_a H_0)^2 - 2^{\frac{1}{2}} c_0 b K T_a H_0]^{\frac{1}{2}}}. \tag{108}$$

Substituting for q from (106) in the preceding,

$$\mathcal{E} = \frac{\pi}{\frac{\pi}{2} + 2\left[2\left(\frac{c_0}{KH_0}\right)^2 + 2\left(\frac{T_a}{b}\right)^2 - 2\frac{2^{\frac{1}{2}} c_0 T_a}{KbH_0}\right]^{\frac{1}{2}}}. \tag{109}$$

Setting $X = T_a/b$ and $P = c_0/KH_0$, and noting that $P = c_0/-4\pi\rho H_0(\partial\sigma/\partial T)$ and $D = 2c_0/AH_0^2 = 2c_0/-4\pi\rho H_0(\partial\sigma/\partial T)$, hence $P = D/2$, where D is given by (87) and (88), we obtain

$$\mathcal{E} = \frac{\pi}{\frac{\pi}{2} + 2\times 2^{\frac{1}{2}}(P^2 + X^2 - 2^{\frac{1}{2}}PX)^{\frac{1}{2}}}. \tag{110}$$

As in our previous discussion, when P is given, $\mathcal{E}$ depends on X.

5.2.1 *Case I*

For $X \gg P$ (or small $b/T_a = \Delta T/2T_a$),

$$\mathcal{E}_{\text{I}} = \frac{\pi}{\frac{\pi}{2} + 2\times 2^{\frac{1}{2}}X} \simeq \frac{\pi}{2\times 2^{\frac{1}{2}}X} = \frac{\pi}{4\times 2^{\frac{1}{2}}}\left(\frac{\Delta T}{T_a}\right), \tag{111}$$

or, as in the preceding case (82), $\mathcal{E}_{\text{I}} \simeq 0.55\times$ Carnot efficiency.

5.2.2 *Case II*

For $P \gg 1$, $X = P$, and $b = 2T_a/D$,

$$\mathcal{E}_{\text{II}} = \frac{2\pi}{\pi + 4.30P} \simeq \frac{\pi}{2.15P}. \tag{112}$$

If b increases further, the radical in (110) goes through a minimum and this gives maximum efficiency; this is obtained by the usual differentiations which yield

$$P - \frac{1}{2^{\frac{1}{2}}}X = \frac{X}{2^{\frac{1}{2}}}$$

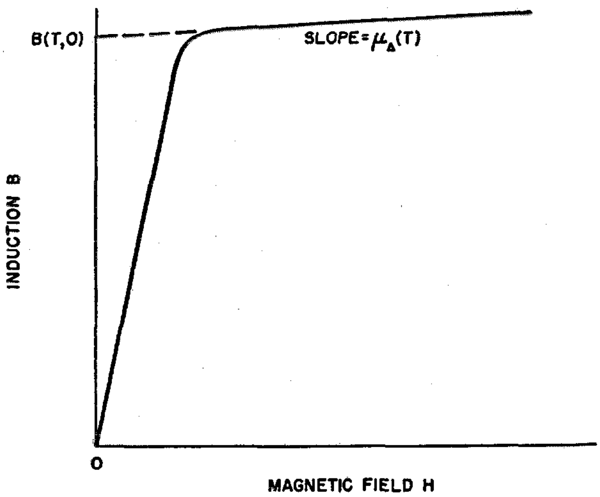


Fig. 8—Magnetization characteristic of the magnetically soft component in a thermomagnetic generator.

and

$$X=\frac{T_a}{b}=\frac{2T_a}{\Delta T}=\frac{P}{2^{\frac{1}{2}}}.$$

Hence,

$$\Delta T=\frac{2\times 2^{\frac{1}{2}}T_a}{P}=\frac{4\times 2^{\frac{1}{2}}T_a}{D}, \tag{113}$$

and

$$\mathcal{E}_{\max}=\frac{\pi}{\frac{\pi}{2}+2\times 2^{\frac{1}{2}}[P^2(1+\frac{1}{2}-1)]^{\frac{1}{2}}}\simeq\frac{\pi}{2P}=\frac{\pi}{D}, \tag{114}$$

when $P\gg 1$.

5.2.3 Case III

For larger values of b, if they can be obtained, we reach a finite efficiency; i.e., for

$$X=T_a/b=2T_a/\Delta T\ll X_{\text{opt}}$$

or

$$\Delta T\gg\frac{2\times 2^{\frac{1}{2}}T_a}{P},$$

$$\mathcal{E}_{\text{III}}=\frac{\pi}{\frac{\pi}{2}+2\times 2^{\frac{1}{2}}P}=\frac{\pi}{\frac{\pi}{2}+2^{\frac{1}{2}}D}\simeq\frac{\pi}{2^{\frac{1}{2}}D}.$$

For the case of Fig. 8, we obtain an efficiency curve that is quite similar to that of Fig. 7, except that the values are increased by a factor of $2^{\frac{1}{2}}$. This is also true of the optimum T (compare (113) and (84)).

Summarizing our results for the two types of magnetization characteristics, i.e., $B=\mu(T)H$ and $B(T,H)=B(T,O)+\mu_\Delta(H)$, we find that the latter yields optimum current and over-all efficiency larger by a factor of $2^{\frac{1}{2}}$ than the former. As already pointed out, the latter magnetization characteristic is also a closer approximation to the actual curve.

6. Increase in Carnot-Cycle Efficiency

The Carnot-cycle efficiency of the proposed system, hence the over-all efficiency, may be increased by connecting a number of these generators in series, using magnetically soft components with descending Curie temperatures.

In such a series arrangement, the heating fluid leaving the first machine would enter the second, then the third, and so on. A similar arrangement could be used for the cooling fluid.

In concluding this discussion, it should be emphasized that the foregoing represents only a first approximation, because the following effects have been ignored:

A. Hysteresis.
B. Nonlinearity in B versus H relations.
C. Nonlinearity in μ versus T relations.
D. Weiss' magnetocaloric effect that corresponds to anomalous specific heat near the Curie point.

These effects would decrease the efficiency and result in harmonics in the output of the generator. It should be added, however, that by a proper choice of the magnetically soft component having a high value of $\partial\sigma/\partial T$, and by using a series arrangement of such generators, as described, it should be possible to improve the over-all efficiency.

Recent Telecommunication Development

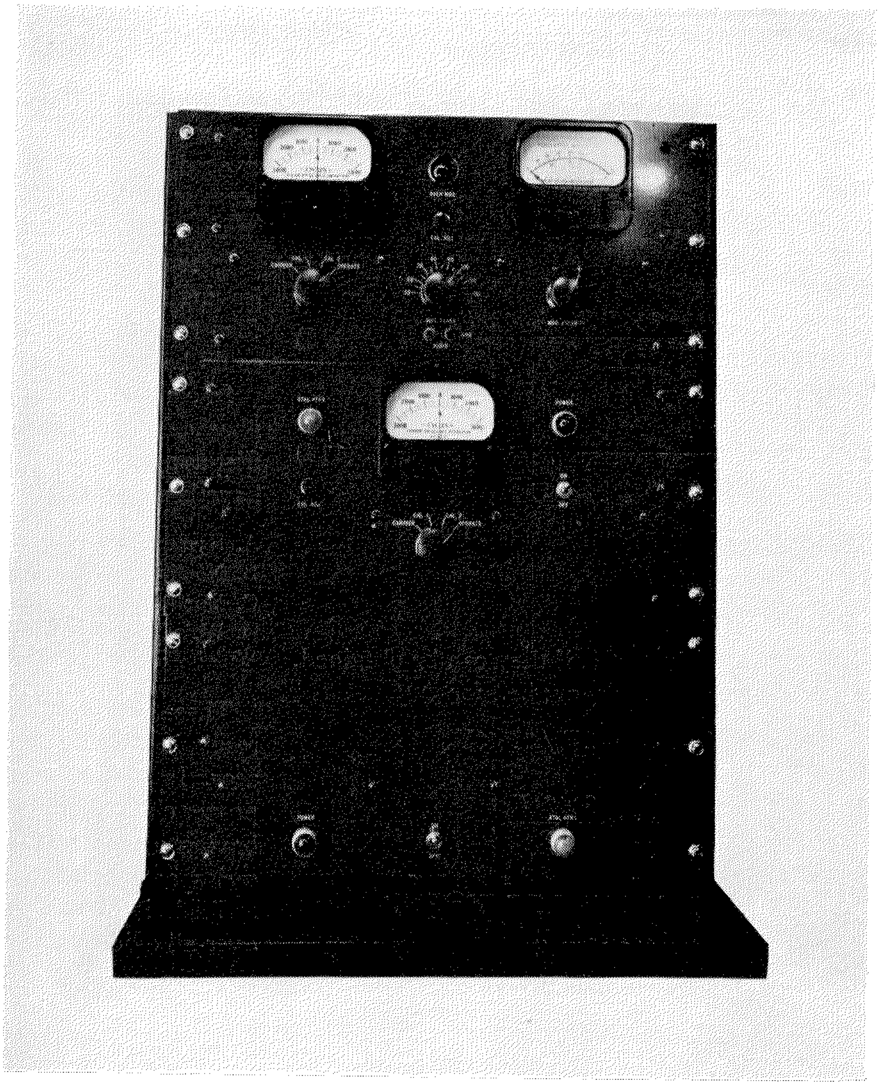
Television Monitor—Federal Telecommunication Laboratories has developed a television station monitor, which measures the carrier frequency of the picture transmission and the center frequency and percentage of modulation of the sound transmission. An overmodulation indicator is provided and will operate an alarm when a preset percentage of modulation is exceeded.

The monitor for the amplitude-modulated picture transmitter has an accuracy of frequency measurement within two parts in a million with proper calibration of the reference oscillator, with a long-time stability of frequency within one part in a hundred thousand. The accuracy of center-frequency indication for the frequency-modulated sound transmitter under conditions of 100-percent modulation is within two parts in a million with proper calibration of the reference oscillator. The long-time stability of frequency measurement is one part in a hundred thousand. The "off-frequency" indicator has a full-scale calibration of ±3000 cycles.

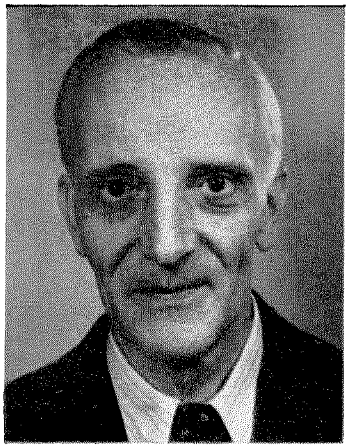
Modulation measurements are based on a frequency deviation of ±25 kilocycles for 100-percent modulation. Modulation percentage and overmodulation indications are performed within ±5 percent.

The sound-transmitter monitor is linear within 0.5 decibel from 50 to 15,000 cycles. The inherent noise level is 75 decibels below a reference level corresponding to 100-percent modulation, and over-all inherent distortion does not exceed 0.25 percent.

Remote indicators may be used for the major monitoring services including the overmodulation indicator and all meters. Monitors are available for any specified channel assigned to television in the U.S.A. (44 to 88 and 174 to 216 megacycles).



Contributors to This Issue



Leon N. Brillouin

Leon N. Brillouin was born on August 7, 1889, in Sevres, France. He was educated at the University of Paris, receiving a doctorate in 1920.

During World War I, he served in the French Army as a radio engineer. In 1928, he was appointed professor of theoretical physics at the Institut H. Poincaré of the University of Paris, and in 1932 joined the faculty of the College de France in the same capacity.

He served as general director of the French National Broadcasting System from July, 1939, to January, 1941, when he came to the U.S.A. From 1941 to 1942, he was a professor at the University of Wisconsin, and spent the next academic year on the staff of Brown University. He was engaged in defense research work at Columbia University from 1942 to 1945, and is now teaching at Harvard University.

Dr. Brillouin has been a consultant to Le Matériel Téléphonique in Paris since 1936, and to Federal Telecommunication Laboratories since 1941.

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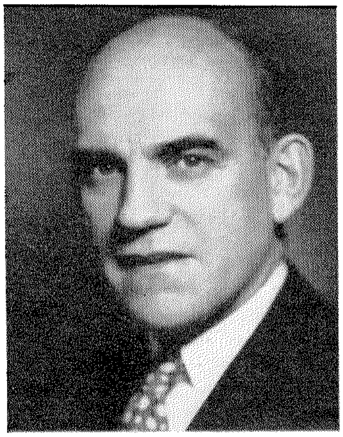

H. H. Buttner

H. H. Buttner was graduated from the University of California in 1915 with the degree of B.S. in electrical engineering. He was employed by the U. S. Navy Department as a civilian employee planning and constructing radio stations on the Pacific coast and in American Samoa. He enlisted in the U. S. Navy and was later transferred to work in connection with the construction of the Lafayette naval transatlantic station at Bordeaux, France. He remained in Bordeaux as a civilian after termination of the war to direct the completion of this station and arrange for its transfer to the French authorities. He returned to the United States as an expert radio aide, U. S. Navy, at the New York Navy Yard.

After serving in the engineering department of the Western Electric Company (later Bell Telephone Laboratories) as a radio development engineer, he joined the International Telephone and Telegraph Corporation in 1926.

From 1930 to 1936, he was Managing Director of International Marine Radio Corporation, an I. T. & T. System associate company.

In 1936, he returned to the United States, joining the group that later became Federal Telecommunication Laboratories. In addition to being President of Federal Telecommunication Laboratories, he is Vice President of International Telephone and Telegraph Corporation, and Vice President and Director of Federal Telephone and Radio Corporation and of Federal Electric Manufacturing Company.

Mr. Buttner is a Fellow of the Institute of Radio Engineers and a member of the Institute of Aeronautical Sciences.

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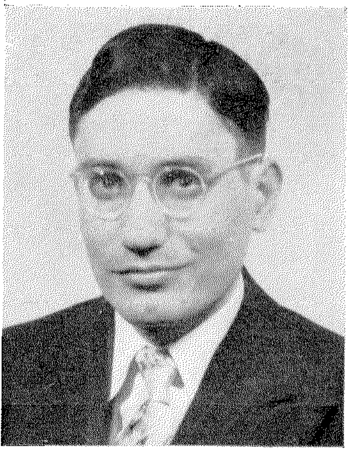

William Dite

William Dite was born in New York City on March 2, 1919. He received the B.E.E. degree from the College of the City of New York in 1940.

From 1940 to 1943, he was with the Signal Corps Laboratories, Fort Monmouth, New Jersey, working on sound-ranging and radar equipment. Since 1943, he has been associated with Federal Telecommunication Laboratories, Nutley, New Jersey, where his work has been largely concerned with communication systems employing pulses.

Mr. Dite is a member of the American Institute of Electrical Engineers.

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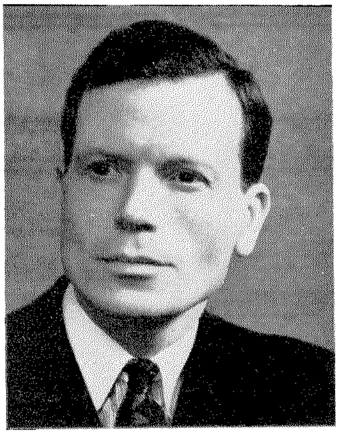
C. W. Earp was born at Cheltenham, Gloucestershire, England, on July 14, 1905. He received the B.A. degree with First Class Honours in 1927 from Cambridge University.

In September, 1927, he joined the International Standard Electric Corporation at New Southgate, and until 1929 assisted in the study of high-frequency transatlantic transmission.



C. W. Earp

Per E. Erikson

In 1929, he joined the Paris laboratory as a radio development engineer where he was chiefly concerned with receiver design, including ship-shore telephone equipments and single-sideband technique.

In 1933, he joined the broadcast receiver development group, which was then transferred from Paris to the Kolster Brandes factory at Sidcup, England, where he became chief of the advance development section.

In 1935, he joined Standard Telephones and Cables at New Southgate as section head for development of radio receivers and direction finders. In 1940, he became head of the newly formed advance development section of the radio division.

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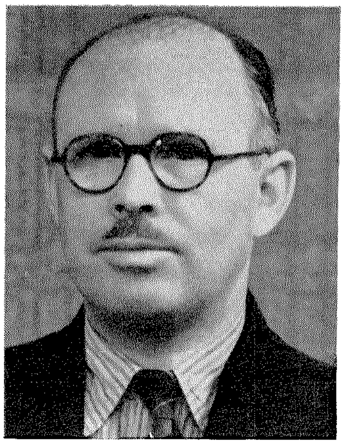

E. C. Fielding

Per E. Erikson received his degree in electrical engineering from the Royal Institute of Technology in Stockholm in 1903, and joined the Western Electric Company in New York as a shop student the same year.

Assigned to the engineering department, he was engaged on the early development of loading coils and balanced toll cables. Appointed transmission engineer for Europe in 1909 with headquarters in London, he was in charge of the construction of the London-Birmingham cable, the first loaded long-distance cable installed in Europe.

During 1918, he carried out the reconstruction of the Rio de Janeiro–San Paulo toll line, the first of its kind in Brazil to be equipped with repeaters and loaded toll entrance cables. As assistant European chief engineer of the International Western Electric Company, he had charge of the design of toll transmission systems, which were being introduced into various European countries after the first World War. In 1928, he was made assistant vice president of the International Standard Electric Corporation and European chief engineer in 1930.

Mr. Erikson has been associated with the Comité Consultatif International Téléphonique since 1925. From 1929 to date, he has been a delegate for the International Telephone and Telegraph System operating companies at meetings of that body and since 1934, he has been secretary of the System's com mittee.

He is a Member of the Institution of Electrical Engineers, London, and a Fellow of the American Institute of Electrical Engineers.

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Haig P. Iskenderian

E. C. Fielding was born in Halifax, Yorkshire, England, on March 14, 1903. He received his technical education at Halifax Municipal Technical College and earned a London University honours degree in mechanical engineering. He was also awarded a royal scholarship tenable at the Imperial College of Science and Technology, London.

He has been with Standard Telephones and Cables in London since 1924, and is now senior mechanical engineer on aircraft radio development and design. He was associated for some years with the technical problems of automatic telephone systems, but transferred his activities early in the war in response to the more important call for aircraft radio equipment for the armed services.

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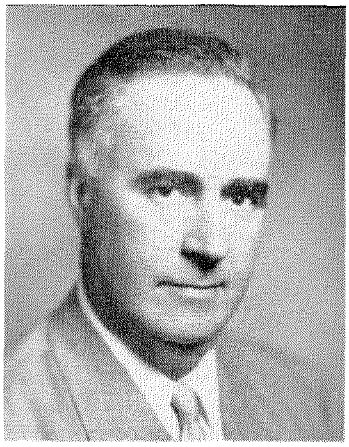
Haig P. Iskenderian was born in Constantinople, Turkey, on January 23, 1905. After completing his secondary education there, he came to the United States in January, 1923, to enter the University of Michigan, from which he received the B.S. degree in electrical engineering in June, 1927.

On graduation, he was associated with the Bell Telephone Laboratories until 1932. He then entered Columbia University and received the Ph.D. degree in physics in 1937. He was an assistant in physics at Columbia University from 1935 to 1937, and an

William C. Lane



E. M. S. McWhirter

instructor of physics at Stevens Institute of Technology during 1937. He then became an instructor of physics and mathematics at Pratt Institute until 1940; also an evening instructor of physics and engineering mechanics at Brooklyn Polytechnic Institute and at Cooper Union, during this period.

Dr. Iskenderian was engaged as a research physicist by the U. S. Navy Department Bureau of Ordnance from 1940 to 1943. He joined Federal Telecommunication Laboratories in 1943, in charge of work on magnetic components.

Dr. Iskenderian is a member of the American Physical Society, Institute of Radio Engineers, and Society of Sigma Xi.

* * *

William C. Lane was born on July 11, 1904, in Goldsboro, North Carolina. He received the B.S. degree in electrical engineering from North Carolina State College in 1928.

From 1928 to 1930, he was with the General Electric Company and during the following three years was employed by Radio Corporation of America, Kolster Radio Corporation, and Federal Telegraph Company. From 1933 to 1936, he was with Hygrade Sylvania Corporation.

In 1936, Mr. Lane joined the engineering staff of Federal Telegraph Company, which later became Federal Telephone and Radio Corporation. In 1947, he was transferred to Federal Telecommunication Laboratories, Incorporated, as a project engineer on military receiver developments.

Mr. Lane is coauthor of the article "Very-High-Frequency Single-Channel Receiver" that appeared in Volume 25, Number 2.

* * *

E. M. S. McWhirter was born on May 24, 1901 in Kent, England. He attended City and Guilds Engineering College of the Imperial College.

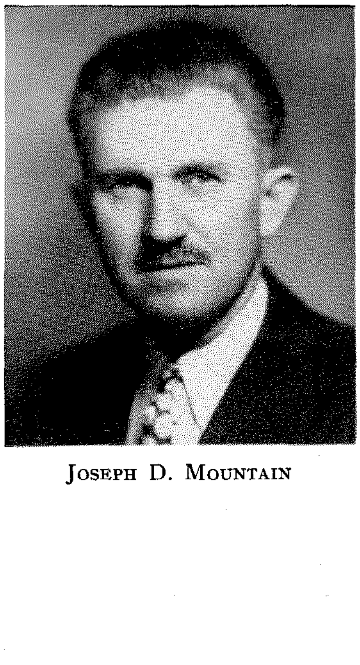
In 1924, he joined the Western Electric Company (later Standard Telephones and Cables) in London as a student engineer. Initially engaged in the design of special apparatus and systems allied to automatic telephone switching practices, he has been closely associated with the development of remote-control systems and became head of the remote-control department in 1935.

He is an Associate of the City and Guilds Institute, and a Member of the Institution of Electrical Engineers. Mr. McWhirter was awarded a Lord Kitchener Memorial Scholarship in 1921 at the Imperial College; and for a paper "Operational Control of Electricity Supply Systems," written in collaboration with Mr. W. Kidd, was awarded the John Snell premium of the Institution of Electrical Engineers in 1946.

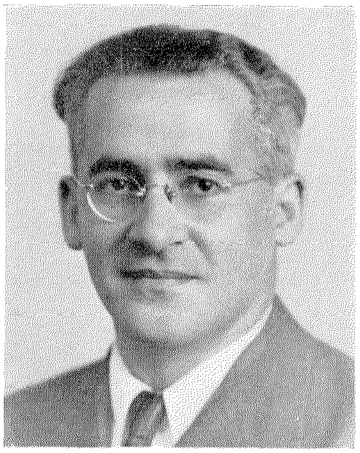
* * *

Joseph D. Mountain was born in Kansas in 1904. A graduate of the U. S. Army Air Service Flying School, he also attended the University of Southern California.

From 1927 to 1937, he was engaged in aerial photography and mapping and spent two years in exploratory expeditions to Saudi Arabia.



Joseph D. Mountain



P. F. Panter

In 1937, he joined Trans World Airlines as a pilot. He served also in several administrative posts including director of training and chairman of the all-weather flying committee.

During World War II, Colonel Mountain was on active duty in Washington and India. He organized and directed the training program of the Army Air Transport Command and in the last year of the war served as executive officer of the air navigation and traffic-control committee of the Secretary of War.

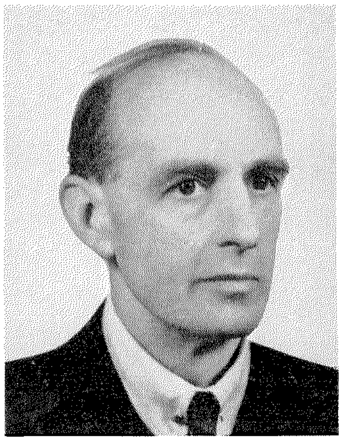
Returning to Trans World Airlines, Colonel Mountain was loaned to Federal Telecommunication Laboratories as a consultant. In 1947, he joined the I.T.&T. System to devote most of his time to the development of Intelex. He is also manager of Federal Telecommunication Laboratories' aviation operation department.

Colonel Mountain holds an army rating of command pilot and a Civil Aeronautics Administration rating of airline transport pilot.

* * *

Philip F. Panter was born in 1908 in Poland. After early schooling in Tel-Aviv, Palestine, he later received from McGill University, Montreal, Canada, the following degrees: B.Sc. in 1933, B. Eng. in electrical engineering in 1935, and Ph.D. in physics in 1936. He continued research in spectroscopy at McGill for an additional year.

After teaching mathematics and physics in Palestine for a year, he



L. C. Pocock

returned to Canada as assistant professor of mathematics and physics in the evening division of Sir George Williams College in Montreal. He served also on the staff of the physics department of McGill University as instructor in physics and later as part-time lecturer, until the end of 1945.

Early in 1941, Dr. Panter joined the transmitter department of the Canadian Marconi Company in Montreal. In October, 1945, he was appointed senior engineer, responsible for the development of frequency-modulation broadcast equipment, at Federal Telephone and Radio Corporation. He later transferred to Federal Telecommunication Laboratories and is now in charge of the theoretical group of the communications division.

Dr. Panter is a member of the Institute of Radio Engineers and the Radio Club of America.

• • •

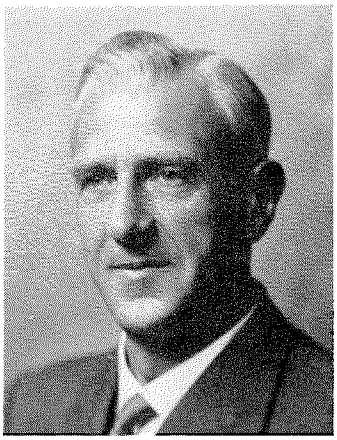
L. C. Pocock received the B.Sc. (Eng.) degree with First Class Honours from City and Guilds of London Institute. He took his M.Sc. (Eng.) degree in 1925, in which year he was awarded a premium by the Institution of Electrical Engineers for his paper on "Faithful Reproduction".

He entered the National Telephone Company a few months before its transfer to the General Post Office at the end of 1911. He then joined the Western Electric Company in London in 1912, and in 1915 took over the general apparatus laboratory work at Woolwich.

During the 1914–1918 war he worked as a radiologist in the Anglo-Russian Hospital in Russia. On his return to the company, he continued in laboratory work, specialising in subscribers' apparatus problems.

He has served for many years on the Comité Consultatif International Téléphonique committee dealing with subscribers' apparatus and transmission measurements, and is chairman of International Telephone and Telegraph System Technical Committee on Electroacoustic Systems.

• • •



C. E. Strong

C. E. Strong was born in Omagh, County Tyrone, Ireland, on November 15, 1898. He received the degrees of B.A. from Trinity College, Dublin, and B.A.I. from the University of Dublin in 1922.

In 1917, he entered the Royal Military Academy, Woolwich, receiving later a commission in the Royal Artillery. He served in France in 1918.

Mr. Strong joined the International Western Electric Company in 1922 and participated in the technical work carried out in England on the establishment of the first transatlantic telephone circuit. Later he was responsible for the development of high-frequency transmitters of the type used for the British empire broadcasting system and for numerous overseas telephone services.

In 1929, he transferred to the Paris laboratories, where he played a leading part in the development of high-power broadcasting equipment. In 1939, he was appointed chief radio engineer of Standard Telephones and Cables, London.

Mr. Strong was appointed an Officer of the Order of the British Empire in 1947. He has for many years been an active member of the Institution of Electrical Engineers and was elected chairman of the Radio Section for the current session. He is also a Senior Member of the Institute of Radio Engineers.

• • •

For the biographies and photographs of Henri Busignies, A. G. Clavier, and Sidney Frankel, see Volume 25, Number 2, pages 205 to 207.

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Compañía Internacional de Radio Boliviana, La Paz, Bolivia

Companhia Radio Internacional do Brasil, Rio de Janeiro, Brazil

Compañía Internacional de Radio, S.A., Santiago, Chile

Radio Corporation of Cuba, Havana, Cuba

Radio Corporation of Porto Rico, Santurce, Puerto Rico[1]

[1] Radiotelephone and Radio Broadcasting services.

Cable and Radiotelegraph Operating Companies

(Controlled by American Cable & Radio Corporation)

The Commercial Cable Company, New York, New York[2]

Mackay Radio and Telegraph Company, New York, New York[3]

All America Cables and Radio, Inc., New York, New York[4]

Sociedad Anónima Radio Argentina, Buenos Aires, Argentina[5]

[2] Cable service. [3] International and Marine Radiotelegraph services.
[4] Cable and Radiotelegraph services. [5] Radiotelegraph service.

Laboratories

International Telecommunication Laboratories, Inc., New York, New York

Federal Telecommunication Laboratories, Inc., Nutley, New Jersey

Standard Telecommunication Laboratories Ltd., London, England

Laboratoire Central de Télécommunications, Paris, France

MR. H-GRAYSON

ELECTRICAL COMMUNICATION

Technical Journal of the
International Telephone and Telegraph Corporation
and Associate Companies

INVENTOR OF THE CATHODE-RAY TUBE

THE STORY OF COPPERWELD

TRENDS IN MILITARY ELECTRONIC-EQUIPMENT DESIGN

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SHORT-RANGE POSITION FINDING

DESIGN OF AN IONIZATION MANOMETER TUBE

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EVALUATION OF TRANSMISSION EFFICIENCY

IN MEMORIAM—J. E. KINGSBURY

DECEMBER, 1948

Volume 25 **Number 4**

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A.–B. Standard Radiofabrik plant in Ulvsunda, Sweden. A wide range of radio equipment is manufactured and includes transmitters and receivers for numerous types of service, vacuum tubes for special purposes, and selenium rectifiers.

Ferdinand Braun—Inventor of the Cathode-Ray Tube

By GEORGE LEWIS and F. J. MANN

International Telephone and Telegraph Corporation, New York, New York

IN 1897, Professor Ferdinand Braun, then at the University of Strasbourg, published the first description of the cathode-ray tube.[1] It is characteristic of Braun's whole life and career that the fiftieth anniversary of this invention, said to rank second only in importance to de Forest's audion in the fields of communications and physics, passed without a single public reference to it or to its inventor. This, even though today television is making most rapid strides and the one indispensable component of all television receivers is the cathode-ray tube—or Braun tube as it was originally named.

Cathode-Ray Tube

Braun developed the tube to investigate high-frequency phenomena. Arranged as an oscillograph, this tube provided a means of representing the passage of currents and voltages in respect to time at higher frequencies than were possible with the then-available loop oscillograph, the inertia of whose moving coil limited its upper-frequency range. The cathode-ray tube has proved applicable to all electrical fields in which rapidly occuring phenomena are involved, not only in high-frequency physics, but also in ionospheric research, television, and radar.

The method of producing the peculiar type of rays called "cathode rays" and the possibility of deflecting them by a magnetic field were already well understood in 1897. Long before that time, exhibits known as "physical cabinets" traveled with side shows throughout Europe. They demonstrated the Crookes, Geisler, and Hittorf tubes. Although elementary textbooks of that period mentioned the deflection of cathode rays, no one before Braun had worked out a way of applying this knowledge to a useful purpose.

Besides being able to generate cathode rays and knowing that they could be deflected magnetically, Braun was aware that a cathode-ray beam would cause the walls of the tube, when suitably coated, to glow with flourescent light. He reasoned then that, by concentrating the beam and controlling its deflection, it could be made to draw figures and patterns in light. His tube differed from present-day cathode-ray tubes chiefly in that it employed a cold cathode. His own description[2] of the tube follows.

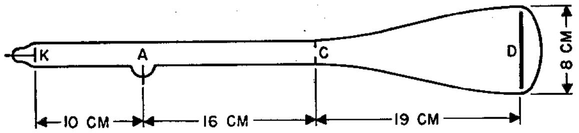
> Fig. 4 indicates the approximate form and size of the tube. The method is based on the deflection of cathode rays by magnetic means (and, if desired, electrostatic means). K denotes the aluminium cathode, A is the anode, C is a diaphragm of aluminium or glass with a centre hole of about 2 mm. diameter, and D is a screen of mica or glass (the latter material is better for photographic purposes) provided with a coat of phosphorescent paint. In order to obtain a smaller speck of light, a second diaphragm may be provided, this diaphragm being placed some 20 cm. in front of C. The back of the cathode K is embedded in

---

[1] F. Braun, "Ueber ein Verfahren zur Demonstration und zum Studium des zeitlichen Verlaufes variabler Ströme," *Annalen der Physik und Chemie*, v. 60, n. 3; 1897.

[2] F. Braun, "Phase-Shifted High-Frequency Oscillations," *The Electrician;* January 19, 1906.

> glass. The cathode ray deflects under the action of the earth's magnetism, but is given the desired direction by a small magnet placed near the tube. In order to deflect the cathode ray by the current under investigation, I provide a small coil (called "indicator" coil) which is placed near to the diaphragm, and whose axis is at right angles to the tube axis. Since the lines of force from this coil are divergent, thus leading to a distortion of the cylindrical flux of cathode rays (and to a speck of light more or less elliptical in shape), it is of advantage to employ two coils, one on the right-hand side and the other on the left-side of the tube. A more homogeneous magnetic field is secured in this way. Such tubes are supplied by Franz Müller (the successor of Dr. Geissler). They do not all turn out equally good. The dimensions given in Fig. 4 refer to one of my first tubes. Since that time the dimensions as also the form of the anode, &c., have undergone changes in a number of ways. When the oscillations are rapid and of a high voltage, the tube is surrounded near the diaphragm by moistened filter paper (or provided with a layer of glycerine) so as to prevent electrostatic effects which may otherwise become very troublesome.



Copy of Fig. 4 from Braun's description of the original cathode-ray tube.

Early Life

Professor Braun was born on June 6, 1850, in the little village of Fulda, principality of Hesse, Germany. Little is known of his early life except that he studied at the Universities of Marburg and Berlin. He left there with Quincke, with whom he was particularly friendly, to become his assistant at Würzburg. Later he became headmaster of the Thomas School in Leipzig. In this position, he also worked extensively in physics. He was one of the few secondary-school teachers of his time who had an enthusiasm for science and the energy to be active in this field despite a strenuous teaching career.

Throughout his lifetime, Braun was noted for his broad interests and simple, even democratic, tastes. He liked to travel and enjoyed painting and sketching. He loved to roam the Tyrolian Alps with his children and invariably he would draw on the backs of penny post cards comical sketches of amusing persons he met on these trips. After he had filled in droll captions for each sketch, he would mail them to his wife or friends back home.

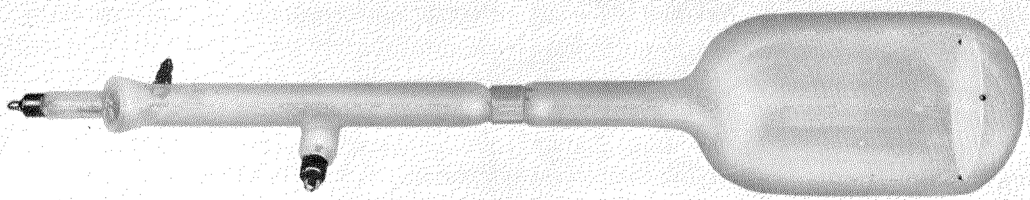
On one such trip, an incident occurred that illustrates the rare and incisive humor that was always present even in classroom and serious conversation but which just bubbled over when he was on a holiday. On this occasion, he visited a small Alpine inn for lunch. Toward the end of the meal, he inquired if there were any chicory in the house. Two unopened packages were brought to him. He then asked if they had an opened package and that, too, was produced. "Now," said the professor with great dignity but with an amused twinkle in his eye, "will you make the coffee for us? I will keep the chicory here to insure that none gets into it."

Although Braun encountered many difficulties in life, his humor never left him. This probably explains his willingness to tackle the most exacting tasks and invariably to triumph over them. For example, he was tone deaf, but he gave brilliant lectures on acoustics.

Academic Rise Rapid

Braun received his first academic professorship in 1876 as professor of theoretical physics at Marburg. In 1880, he went to Strasbourg in the same capacity and then, in 1883, to Karlsruhe. At the time Braun was at Strasbourg, Kundt was working there. The two men had many things in common in relation to their scientific work and Braun was apparently in very good standing with the older man. From Karlsruhe, Braun went to

Birthplace of Braun at Fulda, Germany.



Photograph of an original type of Braun tube. The anode, in this type, was located in the side of the tube. An aluminum thimble located halfway between the cathode and screen provided the aperture that formed the beam. The fluorescent material was supported on a flat mica screen, the observer's side of which was divided into centimeter and millimeter squares. This tube is now part of the historical collection of vacuum tubes of R. McV. Weston.[3]

the University of Tübingen in 1885 as ordinary professor of experimental physics.

At that time, it frequently happened that a private lecturer, who became well known as a result of outstanding experimental work, or as a professor of theoretical physics, or who had published the results of theoretical investigations considered to be outstanding, became ordinary professor of experimental physics.

Braun's appointment to the University of Tübingen was based on his early work in physics. One of his first problems was that of explaining the interactions in galvanic elements. He correctly determined the method of computing electromotive forces from thermic data. Previous to his work, it was believed that the electromotive force of a galvanic element could be computed from the heat of reaction generated in the chemical process. The shortcomings of the old viewpoint were disclosed by Braun through certain theoretical considerations and also by measurements, which are still valid. The electromotive force may be larger or smaller than the value derived from the heat of reaction. To achieve a clear explanation, Braun applied the second law of thermodynamics in a paper that appeared in 1878. In 1882, Helmholtz treated the problem by introducing the idea of free energy.

Braun was also the first to investigate the dependence of pressure on electromotive force in a galvanic chain. This special problem was treated fundamentally by Gans at the Strasbourg Institute. Occupation with these problems stimulated Braun to establish the Braun-Lechatelier principle. The principle did not prove valid in the general form in which it was expressed. However, Braun's work attracted attention because it contained proof that the treatment up to that time was based on an incorrect assumption.

Founds Institute of Physics

Braun's appointment to Tübingen involved a very responsible task for one of his youth, for he was given the assignment of setting up an institute of physics at the university. This assignment brought him, as it would no doubt have brought anyone in the same position, much joy as well as a great many difficulties. However, his efforts proved successful and he found much to enjoy at Tübingen.

Studies Electrocapillary Reactions

In 1891, he discovered a new electrolytic phenomenon he called *Electrostenolyse.* If current is allowed to pass in an electrolyte through a narrow slit, for example, the crack in a glass plate, an electrolytic deposition occurs at this slit like that at an intermediate electrode. This remarkable result was investigated carefully by Braun. In connection with this work and Bequerel's investigation, Braun looked further into electrocapillary reactions. These are reactions between two solutions that diffuse into each other through the pores of a porous wall which separates them.

At about the same time, Braun busied himself with phenomena involved in what were called *Tropfelektroden.* He indicated that Pellat's conclusion that there is no potential difference between the metal and a solution of its salt is incorrect. He gave consideration to the chemical reaction resulting from quicksilver dropping into liquids, or, as it would now be described, adsorption reactions. Further, he was the first who formulated the relation between volume change occurring in dissolving a salt in its saturated solution and the dependence of the solubility on pressure; also, that he placed this relationship on a thermodynamic basis. Braun further proved the interesting fact that compressibility of water in which certain salts are in solution is less than of pure water.

---

[3] R. McV. Weston, "A Museum of Electronic Apparatus," *Electrical Communication*, v. 17, pp. 133–142; October, 1938.

Braun was always primarily concerned with the consideration of principles. However, not only are his assumptions and results significant, but also the methods of his experimentation. He possessed an extraordinary talent and high originality in devising means of facilitating experiments. Consequently, his measuring arrangements and apparatus were of great interest and have been widely adopted in physical investigation and measuring technique. To give an example, in his study of the dielectric constant of rock salt for different crystallographic orientations, he described a simple electrometer. This ingeniously constructed instrument is notable for simplicity, convenience in handling, and low capacitance. Known as the Braun high-tension electrometer, it is still widely used. It is regarded as ideally suited for laboratory demonstrations.

Writes About American Indian

During this same busy period, although he won further honors with his serious scientific work, Braun must have found time to continue pursuit of his hobby of traveling, for he also undertook the task of writing a book about the American Indian. This was a work of juvenile fiction entitled *Conanchet oder Die Ansiedler in Connecticut* published in Leipzig in 1891.

When Braun reluctantly left Tübingen in 1895 for Strasbourg, the Mathematical Society of Tübingen presented him with a farewell gift, which indicated that his good humor and earnestness had won him many sympathetic friends during his decade there. The gift was quite unusual and resulted from his habit, particularly at lectures in physics, of estimating roughly a result in advance or of calculating it, as he called it, "on pad and paper." In fact, the story is told that at one lecture he had to calculate 2 × 25. He immediately set it down on "pad and paper" and went on with his lecture as follows: "Instead of taking 2 × 25, we take 2 × 30, the result should, therefore, be around 50." Fittingly, his farewell gift was a one-place table of logarithms to be used "to calculate on pad and paper."

Wins Nobel Prize

It was when Braun returned to Strasbourg that he began the research on radio communication that won for him, jointly with Marconi, the Nobel Prize awarded in 1909 in physics. Although it is generally remembered that Marconi received the Nobel Prize, few persons today realize that the Italian inventor actually shared this honor equally with the German scientist. Among the reasons for Braun's relative obscurity was his own attitude. He considered himself a scientist and not a publicist. He had no patience with pomp and show.

His democratic, simple attitude toward himself and his work is perhaps best illustrated by an incident that occurred while he was at Strasbourg. At the time, a chair became free at the University of Berlin and Professor Braun was asked to confer with the Chairman of Education. A professorship at this most important institution was then probably the highest honor to which any German could aspire. "Professor Braun, how would you like to come to Berlin and what are your conditions?", he was asked.

His characteristic reply was, "I do not know the value of money in Berlin. Where I am, I can walk downstairs to work and, if I want to work until midnight or until 2 A.M., it is not far to go to my bed. But, if I come to Berlin, I am told I would need a car and a chauffeur. Where I am I can take a third-class railway carriage to the mountains. In Berlin I would need a first-class carriage on an overnight trip for myself and my children. Thus, my conditions are: Can you match Strasbourg—match it in the time and freedom to work—match it in time and freedom to live, and in the atmosphere I love?"

Justification for Braun's share in the 1909 Nobel Prize is probably best given in the following excerpt from the speech addressed to him and Marconi on the occasion of the award.

Marconi's original arrangement had one weak point. The electric waves sent out by the transmitter were of comparatively small power and were composed of waves rapidly decreasing in intensity, so-called damped waves. For this reason, the influence of the waves on the receiving system was only small and the waves of different sending stations could disturb the receiving system. It is principally due to the intelligent and thorough efforts of Professor Braun that these difficulties have been overcome. Professor Braun changed the methods of producing electric waves in such a way that more powerful and less damped oscillations could be emitted. By this method only, was long-distance telegraphy rendered possible, as the waves thus transmitted were such that they could influence the receiving system in the most efficient way by means of resonance. The further advantage was gained by this new arrangement that mainly waves of the frequency of those emitted by the sending station could have an effect on the receiver. The great successes in wireless telegraphy that we have witnessed during the past years could be achieved only by the introduction of this new method.

Early Work in Wireless Telegraphy

About 1898, Braun was approached by members of the Stollwerck family, famous as chocolate manufacturers, for advice regarding an opportunity to invest in a new wireless company. He proved that the company had been formed around a scientific fake. These investigations stimulated his interest in the then new science and eventually led him to the discoveries and improvements credited to him. He invented the so-called closed circuit for radio transmitters and receivers. He also received credit for the substitution of the crystal detector for the clumsy coherer and he carried on pioneering experiments with loop antennas.

At the time when Professor Braun began his work on the problems of wireless telegraphy, popular interest in the subject had reached its peak. As early as 1865, Maxwell had published his paper "On a Dynamical Theory of the Electromagnetic Field" in which he stated the belief that electric action is propagated through free space in the form of a disturbance that travels with the velocity of light. In 1888, Hertz published an account of his experiments on the electromagnetic effects of rapid electrical oscillations and showed that the result of such oscillations was the propagation through space of periodic disturbances that had the characteristics of wave motion. Two years later, in France, Branly used a tube containing iron filings, the coherer, as a detector for Hertzian waves. In 1892, Crookes delivered a lecture predicting long-distance wireless communication using Hertzian waves.

With so many leading scientists approaching and predicting actual transmission of messages without wires, it was natural that interest should grow in intensity. What was also likely from such a situation actually happened; persons with little or no scientific knowledge were eager to put wireless telegraphy on a useful, paying basis. Still, scientists continued to approach the practical realization of their dreams. In 1894, Lodge provided Branly's coherer with an automatic tapping device or decoherer, which improved its usefulness as a detector of Hertzian waves. In 1895, Professor Popoff utilized the coherer with the Lodge "tapper" and a relay, together with an antenna wire and ground, for the detection of lightning and static. About this apparatus, Professor Popoff wrote[4] in December, 1895:

> I entertain the hope that when my apparatus is perfected it will be applicable to the transmission of signals to a distance by means of rapid electric vibrations—when, in fact, a sufficiently powerful generator of these vibrations is discovered.

Scientific Versus Cut-and-Try Experimentation

At about the same time, Marconi was making transmitting tests with a Righi oscillator and attached antenna. For receiving, he utilized an arrangement essentially similar to that of Popoff. As early as 1896, Marconi was actually sending messages with this apparatus over distances of several hundred yards and on New Year's Day of 1898 he transmitted 18 miles to a tugboat from Needles on the Isle of Wight, using a transmitting antenna 120 feet high. A year earlier, Professor Slaby of the Berlin Institute of Technology claimed to have transmitted approximately 34 miles over land using an antenna 300 meters long carried by balloons. However, these experiments were merely of the cut-and-try variety. For example, during one of Marconi's trials, he spent two fruitless days trying to get his apparatus to work. Then he added 20 yards of wire to the antenna and signals were received perfectly. Since Marconi at that time used no tuning apparatus, the additional wire probably brought the receiver closer to resonance with the transmitter.

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[4] J. J. Fahie, "A History of Wireless Telegraphy," 2nd edition, Dodd, Mead and Company, New York, 1902; p. 205.

At this point it was clear that a knowledge of the basic principles involved was essential. Braun in his characteristic, scientific, tireless manner set out to determine these fundamentals. No one at the time knew exactly what was happening. For instance, no one knew the wavelength or oscillating frequency of the waves generated by the Marconi transmitter. It was thought that very short waves of the order of one meter were generated by the spheres of the spark gap in the Righi oscillator.

Braun recognized that this could not be correct. He reasoned that the antenna, which was grounded through the spark gap, should be considered as a whole. The Marconi transmitter, therefore, did not operate with Hertzian waves in the narrow sense, but with much longer wavelengths. This view, which now appears self-evident, was then by no means universally recognized.

It is also of significance that Braun approached the whole problem as a scientist. He seemed to have no patience with the practice prevalent among some contempary researchers of developing radio as one would bake a pie, that is, by adding something here or modifying something there, but without investigating the principles involved or measuring experimental results accurately. Braun belonged to the small group of excellent theoreticians and physicists to whom the really basic and fundamental radio developments and discoveries should be credited. They were tireless, exacting workers who shunned publicity. Consequently, their less-scientific but more-voluble contemporaries often overshadowed them in public recognition. However, the many fine laboratories now in operation throughout the world, some of them shiny with newness, where emphasis on scientific methods has and will continue to produce remarkable advances in radio and electronics, are perhaps more fitting monuments than dubious publicity to the type of leadership Braun represented.

While Braun insisted on measuring his results and checking his theories quantitatively, at the time very little laboratory equipment was available to measure rapid oscillations and feeble currents. Measuring methods for direct or commercial alternating currents were not applicable in the radio-frequency spectrum. In principle, the application of resonant phenomena for measurement applications was known since Bjerknes, but these new methods had to be arranged fundamentally for the newer art. It was precisely to make such measurements that Braun developed the cathode-ray tube.

It is a further tribute to Braun that he surrounded himself with capable men who were able to follow his demands and provide much needed apparatus and help in developing techniques. Men like Cantor and Professor Zenneck, who later became important researchers in their own right, were his able assistants. They enjoyed a sense of freedom and naturalness that went a long way toward instilling the devotion and tirelessness that was even more precious to Professor Braun than the respect he received. No better proof of this can be given than to relate that often Braun would go to the bedroom of one of his assistants, sit down on the bed, and informally discuss the matter in hand. It was not that his assistants were all bedridden by illness, but, in the intensity of their work, they often lost all track of time, working through the night and sleeping in the daytime.

Always most considerate and kind, it seemed the most natural thing to Braun to visit an assistant if he had not shown up in the laboratory. At such times, Braun was prone to intersperse his technical description with jokes that he made up as he unfolded his latest scientific views.

Application of Closed Circuit

At the time of Marconi's work with the Righi oscillator, a means for generating long-wave oscillations—the so-called condenser circuit—had been developed. The oscillatory charges of such circuits had been investigated by Feddersen as early as 1862, and later Lodge made resonance experiments with them. Braun immediately recognized the advantages of applying condenser circuits to radiotelegraphy. He reasoned that the wireless-telegraph transmitter presented two types of problems. First, the strongest possible high-frequency alternating current had to be generated and then this current had to be converted to radiation, i.e., to electromagnetic waves. The Marconi transmitter operated with an open oscillatory circuit, which acted as a good radiator, but which imposed definite limitations on the generation of power. The condenser circuit radiated poorly but permitted strong high-frequency currents to be generated. By combining both, the renowned Braun coupled transmitter was achieved. In its original form, it consisted essentially of the condenser circuit with the spark-gap path driven by an inductor, coupled directly or inductively to an antenna that contained no spark path. As is well known, the oscillations in the condenser circuit generated current in the radiating antenna.

Regarding the importance of Braun's discovery, one can in most cases consider it as characteristic of a basic invention that its importance usually extends far beyond the immediate application for which it was created. This applies to the Braun condenser circuit. The whole technique of radio transmitters has, since Braun's introduction of the closed circuit, seen many variations. The transmitter with the strong spark came first, then the Wien transmitter employing the fundamental discovery of the quenched spark. While in the Marconi spark transmitter, it was possible to use the open circuit, with the Wien transmitter the closed circuit was indispensable. Then came the era of undamped waves with first the Poulsen arc, later the Alexanderson alternator, and finally the vacuum-tube transmitter. All of these later methods of transmission employed the closed circuit almost exclusively.

An analogous observation applies to receivers. Braun not only applied the principle of the closed oscillatory circuit to transmission but he also developed the coupled receiver, which employed the same principles. Methods of reception have also undergone a great variety of modifications in the course of time. The application of the vacuum tube brought about a radical change and provided possibilities that could not have been anticipated in the early period. But, as with the most modern transmitting arrangement, the application of the closed condenser circuit proved to be one of the best and most useful means for achieving the highest possible selectivity.

Thus when, in radio, coupled oscillatory circuits are associated with Braun's name, there is full justification for this identification. For while it is true that the condenser circuit as an electrical oscillatory system was known long before radio communication, it is undoubtedly to Braun's merit that he appreciated its significance and introduced it in a practical form.

Braun's Assistants Develop Wavemeter

Using Braun's ideas, his assistants worked out basic measuring techniques for wireless telegraphy. They led to the first practical Köpsel-Dönitz wavemeter, which consisted of a tuned circuit employing an inductor, calibrated capacitor, and thermic current indicator. To this day, the wavemeter has remained unchanged in principle.

In the same class is the cathode-ray tube. About it, Zenneck writes:[5]

> When Braun brought out his tube, I was very enthusiastic about it. It was exactly what I had wanted for a long time, a device with which one sees what is going on in the current circuit. Later, I considered it a sport to discover as many possibilities of application as possible.

Probably the first and most important contribution Zenneck made to the development of the Braun tube was the introduction of the sawtooth wave that electronically changed the one-dimensional straight-line back-and-forth figure of the cathode-ray spot into a two-dimensional amplitude-time curve. Braun had used a moving mirror to obtain this effect and later the spot was photographed on a film or plate that was moved with uniform velocity perpendicular to the direction of the motion of the spot. Both of the earlier mechanical methods limited the frequency that could be examined, while the Zenneck electronic method was almost limitless. It is, in fact, the same method as is now used in television scanning.

Braun Works with Crystals

As early as 1874, Braun published a report on investigations he had made of certain low-conductivity crystals. He discovered that if he passed an electric current through copper pyrites, sulphur pyrites, galena, "fahlerz," and similar crystals, he found that the strength of the current was not proportional to the electromotive force. If, in addition, the electrodes were constructed differently, the current strength was also dependent on the direction of the applied potential difference. Braun, in his first experiments, for example, encountered differences in strength of 30 percent depending on the direction of current flow. In later arrangements, the difference in current strength was still greater so that, practically speaking, it could be said that the current was conducted in only one direction. This deviation from Ohm's law is designated as unipolar conductivity or rectification.

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[5] J. Zenneck, "To the Fiftieth Anniversary of the Braun Tube," Unpublished.

Braun found an early application for it in the crystal detector, which he first used to receive wireless telegraph signals in 1899.

Before the crystal detector, the coherer was the only means of detecting radio waves. This remained an exceptionally critical and inconstant piece of apparatus not really satisfactory for practical use, despite the efforts and improvements of a number of inventors. Many investigators, therefore, earnestly sought a better detector. As with the closed circuit and the cathode-ray tube, Braun's crystal detector, which for simplicity remains unexampled today, achieved notable results. For a long time in fact, it replaced all other detectors and completely dominated this field until the vacuum tube was introduced. In recent applications of radar and ultra-high-frequency operation, there has been a return to the use of the crystal detector as a simpler and even more efficient device than the vacuum tube. Also, Braun's originally discovered principle has seen wide application in metallic rectifiers for power frequencies. A recent development, the transistor, applies the principles of semiconduction to the amplification process, thus completing the circle. The vacuum tube supplanted the crystal detector. Then many new applications were found for the vacuum tube. Now it is predicted that the transistor, which is little else than a more advanced crystal detector, will supplant the vacuum tube for many of its applications, including amplification and oscillation.

Experiments with Directional Radio

Another field pioneered by Braun was that of directional transmission and reception. As early as 1902, he made successful trials with an antenna inclined about 10 degrees from the horizontal, showing it to be more sensitive to waves received from a single direction. From this type, a specific class of receiving antennas was constructed and included the inverted-L and the horizontal or earthed antennas. Similar antennas were proposed for transmission, but for this purpose their directivity proved very low.

Braun attacked the problem of directive transmission differently from that of reception. The principle of his method was to employ two or more separated vertical wires. In the initial experiments, he employed three wires set at the corners of an equilateral triangle.[6] The wires were supplied with currents differing in phase. Phase displacement and separation distances were chosen so that the wave was beamed in a desired direction. Such trials were made in Strasbourg in 1904. Field strength was measured with a bolometer. Results were as had been predicted theoretically. It is, of course, commonplace now to use such methods of directional transmission, particularly for the higher frequencies.

In 1913, Braun again took up the problem of directive reception. At this time he experimented with loop antennas and was, no doubt, prompted to employ this method of directional reception by the fact that the vacuum tube then offered a more sensitive means of detection than the crystal detector and could also be employed as an amplifier. The advantages of directional reception such as reduction of interference, a greater signal-to-noise ratio, and direction indication were recognized as possibilities by Braun, although their full utilization did not come until later. During these experiments, which Braun conducted in Strasbourg with signals received from the Eiffel Tower, field strength of radio waves was measured to an absolute value for the first time. Today field-strength knowledge at a receiving point is considered indispensable and all modern methods of measurement are based essentially on the early Braun arrangements.

Visits America

When World War I broke out, a patent suit was brought against the Sayville, Long Island, radio station, which worked with a station at Nauen, near Berlin. The purpose of this suit was to close the Sayville station and thus cut off all means of communication between Germany and the U. S. A., the cable between these two countries having already been cut. Braun was asked to appear in New York to testify regarding experiments he had made. At the age of 64, with undermined health, he made the trip in midwinter through the British blockade, because he hoped in this way to be of service to his country.

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[6] F. Braun, "On Directed Wireless Telegraphy," *The Electrician;* May 25 and June 1, 1906.

The suit was indefinitely postponed and Professor Braun stayed on to live at his son's home in Brooklyn, New York. There were no means by which he could return to his laboratory in Strasbourg and later, when the U. S. A. entered the war, he would not have been permitted to return. He was suffering from an incurable illness and was, perhaps, saddened by the fact that the great nations, all of which he seemed to love, were at war. At least, he had shown enthusiasm over the New World after he had returned to Germany from a trip made in 1909 across North America as far as Vancouver, British Columbia. At that time, he gave an illustrated lecture on his impressions of America. He even built a miniature shoe factory that operated on a high-speed conveyor-belt plan just as he had seen on his trip. He fed in small strips of leather belt at one end and, pronto, out came miniature shoes. To illustrate the wonders of Yellowstone Park, he made a model of the area in contour and color. Even "Old Faithful" was represented in the form of a geyser of carbonated water.

Now these happy memories were far in the past. Not only was Professor Braun very ill, but, for the first time, he was forced to be away from his laboratory for an extended period of time. A quick holiday to the mountains was one thing. A jaunty vacation across the Atlantic had only left him with renewed vigor to return to his experiments. But a time had now come when he was confronted with months and months of idleness, an ill man. Even then he filled up some of this time by giving piano lessons to children who could come to his Brooklyn home. And it was while shoveling snow that he suffered the accident that finally caused his death. He fell on the ice and broke his hip. For weeks he lay in bed in pain and then complications set in. He died April 20, 1918. His body was cremated in New Jersey and, after the war was over, his ashes were returned to his native Fulda and buried next to the grave of his parents.

Tributes to Braun

Few men of science have achieved as much as Ferdinand Braun. Also, rarely has a man who has accomplished so many great things also been so loved and respected. He possessed a naturalness which, along with his friendliness and exceptional good will, made him extremely well liked among his associates.

No more fitting tribute to this quality of Braun can be given than to quote the following statements by men who worked with him and knew him well. The first was written by Mandelstam and Papalexi.[7]

> As a teacher, Braun is unforgettable to all who were privileged to work in his institute. He left it to each one according to his individuality to follow his inclination. He followed all work constantly with interest and was always ready with advice and help. And his advice was more than clarifying: Braun had the vision and sensitivity of an experimenter that cannot be learned and with which only the chosen few natural researchers are gifted. These qualities to which, to a large degree, he owes his success as a researcher benefited also his students. . . . Braun's capabilities as a research worker assure him a place of high honor in the realm of science, and all who had the good fortune to come in close contact with him will hold the memory of this great, experienced, intelligent, and good man in affection and the highest esteem.

More recently, Braun's former assistant, Professor Zenneck, now President of the Deutsche Museum in Munich, and recognized as a distinguished scientist and teacher in his own right, wrote[4] of his former mentor.

> Braun was a distinguished experimenter, who had thorough mastery of the experimental aids in physics. In his investigations, he always applied himself only to essentials. He had a fine instinct for what was essential in each case. A good example of this is that he undertook the experiments with the frame antenna right at the time when the amplifying tube entered the field and a widened application of this form of antenna was made possible. . . . Personally, Braun by nature was beyond reproach. . . . Despite all the difficulties which he encountered in life, his humor never left him. His witty remarks at lectures and in conversation were all the more effective because one could see his own enjoyment of them. As Director of the Institute, he was beloved both by the candidates for doctorates and by the assistants. In all, we assistants had in him a chief who advanced our scientific work in every respect and who regarded our individualities with kindly understanding. . . . Braun was what we would call a splendid man. I am convinced that all who had to work closely with him, remember him not only with respect but also with pleasure.

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[7] Von L. Mandelstam and N. Papalexi, "Ferdinand Braun—In Memoriam," *Die Naturwissenchaften;* August 10, 1928.

The Story of Copperweld

THIS is a story about wire—not ordinary wire. It is a story about a comparatively new and different wire, a wire that has successfully joined two great basic metals in a perfect union to carry power and man's voice to places all over the world.

This is also a story about the company that makes that wire. Not a large company, as industrial giants go, nor an old one in the sense of century-old enterprises. A company whose people for 33 years have lived and worked with one primary aim—to develop and improve a product that others had unsuccessfully tried for decades to develop.

The company began operations in 1915 with 23 workers. It now numbers its employes in the thousands and its product spans the continents of the earth. The product is a special kind of wire called "Copperweld" and the company is the Copperweld Steel Company.

1. The Art of Making Wire

The story begins at the dawn of history, some 3700 years ago. Making wire for ornaments, tools, hooks, fasteners, for a score of other uses in a primitive world, was one of the most ancient of metal-working arts. The wire maker was one of the most honored of metal craftsmen. Gold wire, a Pharoah's necklace, is mentioned in the Bible. Samples of Assyrian and Babylonian wire have been found dating back to 1700 B.C. Almost nothing is known of this early manufacture, but the wire was probably shaped with hammers.

Two thousand years ago man made a discovery that was, in its field, comparable to the discovery of the principle of the lever, to the invention of the wheel and the windlass. He found that a hammered brass rod could be made smaller and smaller by being pulled through a tapered hole in an iron plate. Thus the principle of the draw plate was born and the first wire was drawn.

A Latin manuscript written by Theophilis sometime in the 8th or 9th century describes a method of making wire of an alloy of lead and tin. The alloy was cast into an ingot and then hammered into a long bar. The pointed end of the bar was forcibly drawn through a series of holes set in a draw plate. In this way, the cross section of the rod was gradually reduced to that of the last hole through which it was drawn, its length being correspondingly greatly increased.

Some scholars believe that the chain armor used by the knights in the great crusades against the Saracens and the Turks was made of drawn, rather than hammered, wire.

The earliest known mention of commercial wire drawing was in 1351 and again in 1360. In that time, in the now-ruined cities of Nuremberg and Augsburg in Germany, the wire *smith* was joined by the wire *miller*. In this early manufacture, the wire miller attached the end of the wire to a belt around his waist and pulled it through the draw plate, a few inches at a time, by stepping and falling backwards. A windlass type of pull was introduced a few years later, followed by an application of water power.

Until 1565, English iron wire was drawn by hand only, with such poor results that almost all wire was imported from Germany and Holland. The first wire mill in England was set up by Dutch manufacturers in 1662.

A "wyer drawer" named Nathaniel Robinson, who set up a mill at Lynn, Massachusetts, four years later, is the ancestor of the American wire industry. But 168 years later, in 1834, there were only three wire mills in the 24 states of the union, with a total output that year of 15 tons. Wire, which was later to be known as "the product with 1500 uses," was still being applied to but a few basic needs.

Then, within 40 years, two things happened that revolutionized industries and changed the face of America. Wire changed from a product of few uses to become the thong and tendon that bound the land together.

In 1844, S. F. B. Morse built a telegraph line from Baltimore to Washington, a distance of 40 miles, and then laid down a line to connect those two cities with New York. Seven years later more than 50 companies were operating under Morse's telegraphic patents. In that same decade, a cable was laid across the bed of the Atlantic Ocean.

On March 10, 1876, Alexander Graham Bell spoke a few words, which were to affect every man, woman, and child in the land. To his assistant, who was listening in the next room at the end of a wire, he said, "Mr. Watson, come here, I want you."

By 1878, New Haven had a telephone switchboard with 21 subscribers. By 1920, America had spun a wire network that encompassed 20 million telephones.

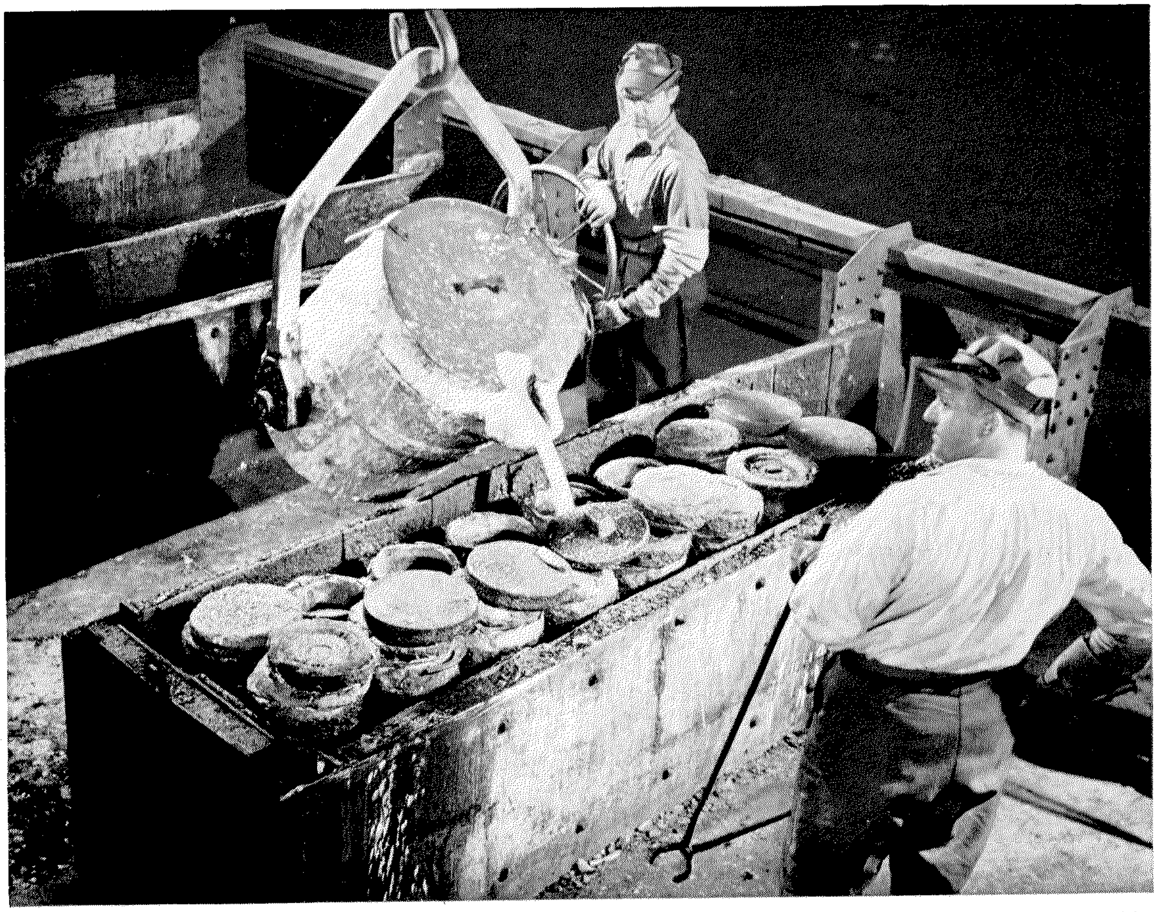
In the meantime, in 1890, the first power station went up. Distribution of electric power was becoming a major U.S.A. industry. Electric power was being carried up and down the land.

These new industries needed wire, millions of miles of it. Much of that wire had to be copper. This brilliant metal had been used by the human race to make its first weapons in remotest antiquity. It was the metal which, joined with tin to form man's first metallic compound, had given a name to the Bronze Age. Now this metal had been found to be the best commercial conductor of electrical energy.

But as distances increased, industry needed and hoped for a special kind of wire that would have the one thing copper lacked, strength. To carry electrical energy, it needed a wire strong enough to withstand great tensions, and which would not corrode. The wire must carry power and voices many hundreds of miles. It must span great rivers and valleys.

2. *Search for a Perfect Metal*

The metal which would make such a wire simply did not exist. Steel was strong enough. But steel was not a good enough conductor, and it corroded. Copper was an excellent conductor and did not corrode. But copper was not strong enough. It had no backbone.

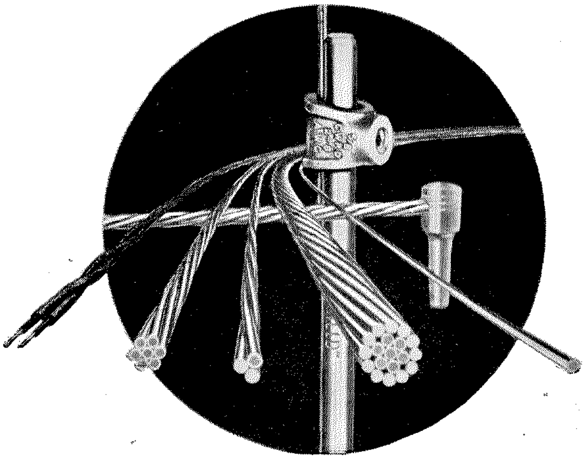


Basic to the Copperweld process is the pouring of molten copper around a heated alloy steel billet centered in a mold. The resulting ingot is reduced by hot rolling and cold drawing to final wire size.

Could the metallurgist combine the two into a wire that would be strong, durable, and conductive? If he could, time and money might be saved in power and communication lines. Distances might be spanned that were impossible for copper wire. The space between supporting structures might be increased from a few hundred feet to a mile or more.

Coating iron with copper was by no means a new idea. The ancient Swiss lake dwellers and certain tribes in Assyria, we now know, had made tools and weapons with an iron core and a copper-alloy covering. An English metallurgist had patented a copper-iron combination in 1821. He melted copper in a shallow cast-iron vessel and applied it to iron pans. His pans were tough, and the copper lining kept them from corroding.

Many men tried to create a steel and copper wire. The first attempt was made in 1860, when two American metallurgists wrapped steel wire with copper strips and then tinned the result. They failed.



Samples of Copperweld products. The cables from left to right are telephone drop wire, strand, two types of power conductors, and telephone line wire. A ground rod with connecting wire clamped at the top is the vertical assembly and the horizontal cable with a plug-type terminal is a railroad signal bond.

Various methods of merging steel and copper by casting followed. In 1885, a Frenchman made a composite billet which could, by careful handling, be rolled and drawn into wire. But the two metals were not joined by a weld; the copper served only as a shell over the steel core. Electrolysis and destruction resulted.

An electroplating process was developed, tried, and discarded because the copper-plated wire rusted.

The Duplex Metals Company, organized in 1905, found the key. That was to *weld* steel and copper together. Their method, the Monnot process, consisted of dipping an especially prepared steel billet into a copper bath, thus *wetting* the steel with the copper. The *wetted* steel billet was then placed in a mold containing molten copper to form a copper-clad ingot.

Five American men with vision then joined forces to take the next big step. Founding a firm named the Copper Clad Steel Company, they invented what is known as the Copperweld process of welding steel and copper. This process, less complex than previous methods, was based primarily on obtaining a *molten weld* between copper and steel.

Steel and copper were now at last merged into a weld so strong that they could not be separated except by destruction of the bimetallic material itself. The next step was to develop processes by which this material might be drawn into bimetallic wire. That was no simple matter. There is a wide difference in the melting and rolling temperatures of these ferrous and nonferrous metals. There are differences in malleability and in temperature coefficients of expansion.

In 1915, the company, now renamed the Copperweld Steel Company, and its 23 employees began the first actual production of a usable wire with a tough steel core and a uniform rust-resisting conductive copper covering. That first wire was found to be satisfactory in every respect.

Research and development of skills over the past three decades have enabled metallurgists to improve the manufacture of bimetallic wire, to lower its cost, and to speed up and simplify its production. Steel and copper can now be maintained in exact proportions down to wire thinner than a hair. And yet the original process was so sound fundamentally that it has remained the basis for any product improvement today.

3. *Copperweld Process*

The process that welds copper to steel and produces bimetallic wire takes place in a half dozen major steps.

Alloy steel billets, about six inches in diameter and four feet long, are picked up by a magnetic crane, cleaned and inserted exactly in the center of a cylindrical graphite mold, which is approximately eight inches in inside diameter. The mold is then placed in a furnace. When the steel billet inside it has reached welding temperature, the mold is removed from the furnace.

Molten electrolytic copper, heated to the proper temperature, is poured into the space between the steel billet and the inner mold wall. It is at this moment that a molten weld is produced between the two metals. Actually, the copper grains diffuse between and into the steel grains, interlocking and joining the two metals into an unbreakable uniform weld.

This weld, a form of iron-copper alloy which is much stronger than copper alone, is the basic factor that permits exacting fabrication. Without the successful weld, the billets would fail during the rolling process that follows. The finished product, moreover, would not be able to withstand the tensional and torsional stresses that a conductor receives in service on overhead lines.

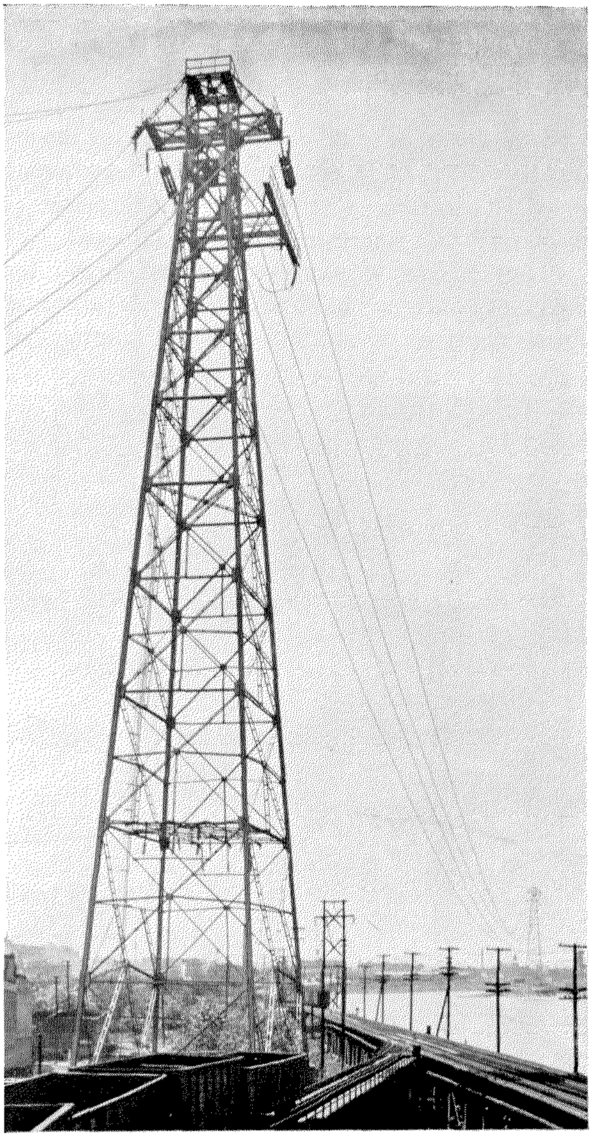
When the copper has been poured, the molds are withdrawn from the pouring bay and allowed to cool. The billets are removed and placed in a preheating furnace and heated to a rolling temperature. That rolling presents a unique problem. Since copper melts at approximately 2000 degrees Fahrenheit, the billet must be rolled at a temperature well below the standard rolling heat for steel. Specially designed rolls reduce the billet without destroying the properties of copper and steel and maintain the original proportions and concentricity of the two metals.

The first rolling operation takes place on a 24-inch water-cooled tilting-table mill. An operator passes the billet between the rolls. As it reaches the opposite side, the table tilts and a second operator guides the billet between another set of rolls. This operation continues until the billet has been reduced to a bar with a diameter of from two to three inches.

Since the bar has become considerably longer in this rolling, it is cut in half. It is then reheated and passed through a series of smaller rolls to be formed into a rod, which is finally reduced to a diameter of three-eights of an inch and coiled.

Next step is to transform the rod from its hot-rolled structure to a fine-grained molecular arrangement, a structure that is important in the product's use as an overhead conductor. The rod is fed from payout reels into two sectional heat-treating furnaces. It is heated above its so-called critical temperature and quenched in water.

The rod is then cleaned in a mild sulphuric-



Copperweld conductors, installed in 1927, are providing dependable service on this 3300-foot river crossing.

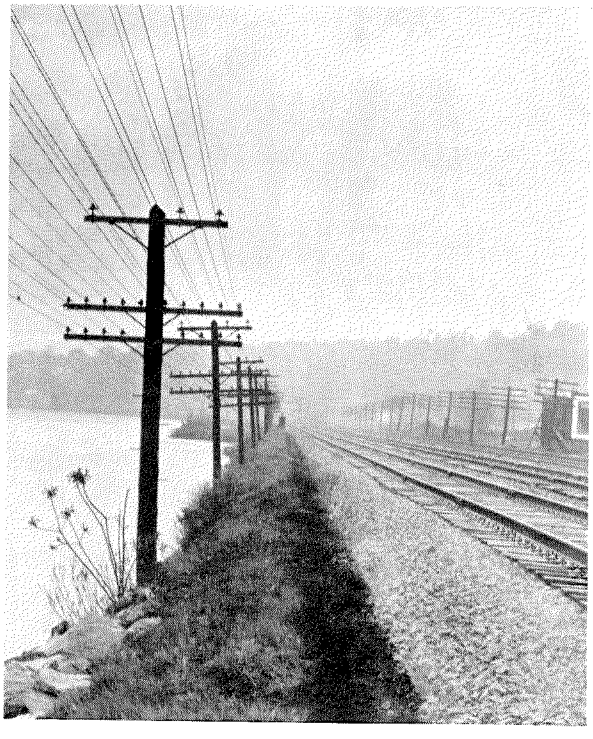
acid solution, which dissolves the copper oxide but does not affect the copper. It is now ready to be cold drawn into wire.

The drawing process has two objectives. It reduces the wire to a practical size for use either as a single or a multistrand conductor; and it increases the tensile strength by cold working.

The cold-drawing operation involves special techniques, yet generally is similar to methods used in drawing copper or steel individually. In particular, the slope and curvature of the dies and the speed of drawing must be closely controlled. The operation begins on huge seven-die wire-drawing machines.

In multiple drawing, the wire is pulled through a single die onto a drum. From here it passes over an idler, through a second die, and onto a second drum. A number of such drawing stages are provided on a single machine. Equipment used in drawing wire to the smaller sizes is essentially the same, differing only in detail.

The finished wire may be coiled singly for shipment or it may be stranded. Wire is produced in as many as 37 strands. Smallest commercial Copperweld wire normally made is Number 20 (0.032 inch), though it has been successfully drawn down to two thousandths of an inch.



Along the right-of-way of a leading railroad where Copperweld plays an important part in the operation of the signal system.

Extensive improvements in strength have taken place in the past three decades. The earlier wire had a tensile strength of 80 to 100 thousand pounds per square inch. In 1920, by use of improved steels, wires were produced having tensile strengths ranging from 120 to 140 thousand pounds per square inch. About 1930, the development of special alloy steels made it possible to manufacture wire having strengths up to 180,000 pounds per square inch.

Improvements in the art of rolling make possible a guarantee of minimum copper thickness as well as a guarantee of minimum conductivity. "Thirty-percent" wire has an electrical conductivity equal to 30 percent of the conductivity of a solid copper wire of the same diameter, with a minimum copper thickness of 10 percent of the radius of the finished wire. "Forty-percent" wire has an electrical conductivity not less than 40 percent of a solid copper wire of the same diameter, with a minimum copper thickness of 12.5 percent of the radius of the finished wire.

Along with improvements of the wire, special tests and inspection practices have been developed and cover such properties as torsion, weld, conductivity, surface, breaking strength, and elongation. These tests are continuous from ingot to finished wire. Methods and equipment have been devised for measuring the copper thickness at any point in a coil without destruction of the wire.

4. *Copperweld Wire at Work*

Copperweld wire varies in size and strength depending on the use for which it is intended. These wires are widely used by electric light and power companies for overhead ground wires on transmission lines, for pole guys, and as messengers for supporting aerial power cables. They are used in various combinations with copper wires as power conductors.

The telephone industry uses this wire extensively as "drop wire" for connecting the subscribers' premises to the company lines, for line wire on open construction, for guying, and for aerial cable messenger.

Railroads use Copperweld signal bonds to complete the electrical circuit from rail to rail. They use signal line wire, guy and messenger strand, and overhead ground wire. Electrified roads use Copperweld-Copper catenary messenger as a combination conductor and trolley-wire support.

Similar composite Copperweld-Copper conductors are used for extremely long spans of transmission lines crossing rivers and valleys, often with a distance of a mile or more between supporting towers.

With the advent of rural electrification in the United States, there was developed a high-strength, low-cost conductor made of one non-rusting Copperweld "strength" wire and two copper "conductor" wires in the form of a three-wire strand. Hundreds of thousands of miles of these conductors in service have been a continuing factor in reducing the cost of rural lines.

Copperweld ground rods and clamps are used in all of the above fields. The corrosion-resisting and easy driving properties of the ground rods make them almost a "must" in every type of electrical work where a driven ground is needed to provide positive grounding protection.

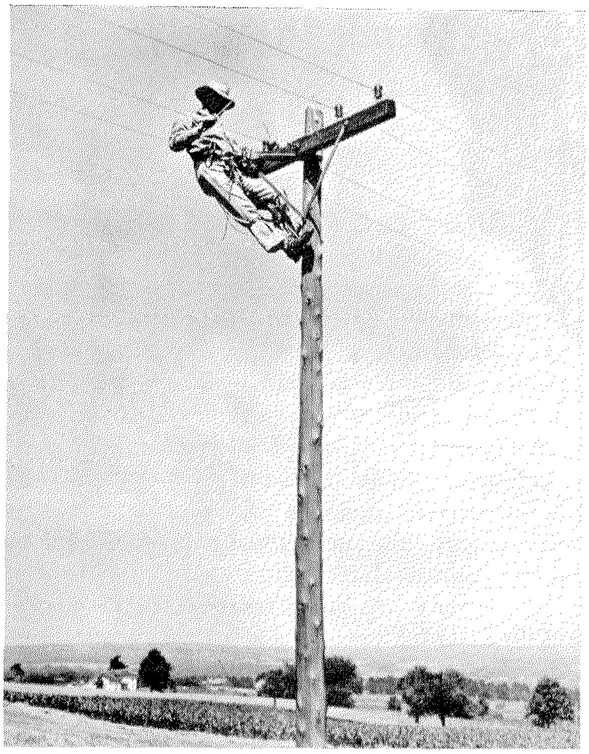
Such are some of the uses of Copperweld wire in the company's primary fields of application: utilities, railroads, and communications. Many new uses have developed in the past few years.

Perhaps the most interesting of these is a unique construction application being employed by the U. S. Army engineers on various flood-control projects. Here Copperweld's permanent strength and corrosion resistance provided the solution to the serious problem of reinforcing-wire failure in concrete revetment mattresses for levee protection.

Reinforcing wire previously used in these installations eventually failed because of corrosive action of the water. Result was that the concrete forms would crack and sometimes give way against the force of the current.

After exhaustive tests, Copperweld proved entirely resistant to the corrosive action and has since been specified for use in reinforcing and tying together articulated concrete flood- and erosion-control mats.

Fabrication of wire into "revetment mats" consists of a series of bending, welding, and stapling operations. These are made so that the wire reinforcement network can be extended in any direction by stapling.



Lineman installing a Grip-Flax tie on a Copperweld telephone wire.

Throughout World War II, Copperweld wire served in every theater and in every battle zone. The U. S. Army Signal Corps, in fact, strung enough Copperweld telephone wire across the battlefields of Europe and the South Pacific to encircle the world 20 times. Along the Alcan highway, enduring some of the world's severest weather, more than 10,000 miles of Copperweld telephone wires have been carrying the weight of wind and ice since 1942.

Such is the story of Copperweld wire and of the company that developed it and makes it. As the network of power and communication systems grows greater and greater, Copperweld engineers envision new horizons for a product that has been an integral part of the electrical industry for a third of a century. Where the basic requirements are the combination of strength, durability, electrical conductivity, and rust resistance, Copperweld has an indisputable performance record. It is pledged to continue and expand that record.

Trends in Military Electronic-Equipment Design

By NORMAN H. YOUNG

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THE SPECIFICATIONS furnished to development laboratories by the armed services as part of their procurement program display a steadily varying pattern as the needs of the services and available techniques to meet these needs change. These variations sometimes reflect the requirements of an individual tactical problem, but on examination of many specifications a general pattern of growth toward certain goals is discernible. For this achievement, the government is prepared to expend considerable effort.

While some of these targets are of purely military usefulness, others reflect commercial considerations of great importance, and their significance will increase as techniques are improved under the stimulus of the military development program. All engineers, therefore, whether engaged in government-sponsored programs or not, should examine their activities with care to see that these trends are grasped as quickly as possible and turned to use wherever they fit the program.

1. *Detailed Analysis of Trends*

One common spirit pervades the entire development program. In essence, it is the heart of all technological improvement. This is best expressed as a vigorous impatience with all the inconveniences and shortcomings of the apparatus at hand and the endless repetition of the question: "Why isn't it perfect?" All the characteristics to be considered arise from the desire to have electronic servants of zero size and weight, responding with infallible accuracy under all conditions imaginable. The degree to which an engineer refuses to accept *any* limitation as necessary or incapable of improvement, is directly a measure of his conformity to the spirit of this program.

1.1 MINIATURIZATION

Military operations invariably place the maximum burden possible on available means of transport. Space and weight are equally at a premium on the person of an advance scout or in the hold of a supply vessel. Therefore, components are being developed that open new possibilities in the way of space and weight reduction. The trend from so-called "standard" tubes to "miniature" types is well advanced in commercial design, and in military apparatus the use of "subminiatures" should always be reviewed in the light of the specific problem at hand.

Miniaturization brings with it additional problems of heat dissipation for it is not always possible to decrease the power handled by the apparatus. These difficulties may be overcome in some cases by the application of special components and materials, permitting a much greater rise in operating temperature over the surroundings than has been generally used in the past. For example, the use of glass insulation in transformers has permitted unusually small units to operate at a much higher temperature than was possible in former designs.

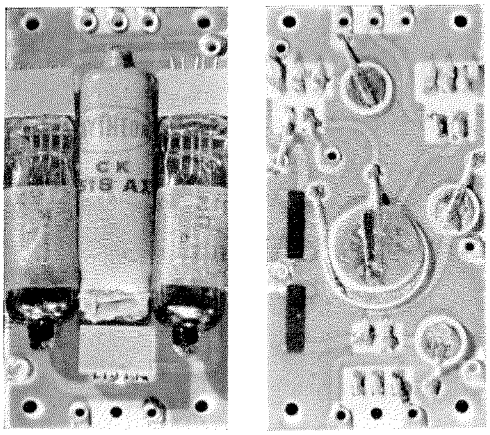
1.2 UNITIZED CONSTRUCTION

Unitized construction hits at another facet of the problems in military equipment. Electronic apparatus that fails in the field must either be so easy to repair that relatively untrained personnel can restore it to a guaranteed level of performance, or must be capable of being completely replaced; the defective units either being discarded or returned to a major base for repair and adjustment. In particular, it must be impossible for the operator partially to repair an equipment so that it apparently operates, but is actually presenting erroneous information.

The only equipment that can be serviced in the fashion outlined is one in which each major component or subassembly is complete in a self-contained unit, usually "plug-in" and hermetically sealed. The sockets for a group of such plug-in assemblies can be made sufficiently rugged to insure very great dependability. Field servicing is then reduced to interchanging spare units for defective equipment until normal operation is restored.

1.3 Increased Reliability

While the need for absolutely dependable equipment has always guided designers of military apparatus, one important change in techniques throws special emphasis on this problem. The increasing use of multiplexing to carry the heavy burden of military traffic means that the failure of one component in certain parts of an equipment may disable not one circuit, but dozens, or in the future, even hundreds of circuits. The use of unattended repeaters also imposes additional requirements of reliability.



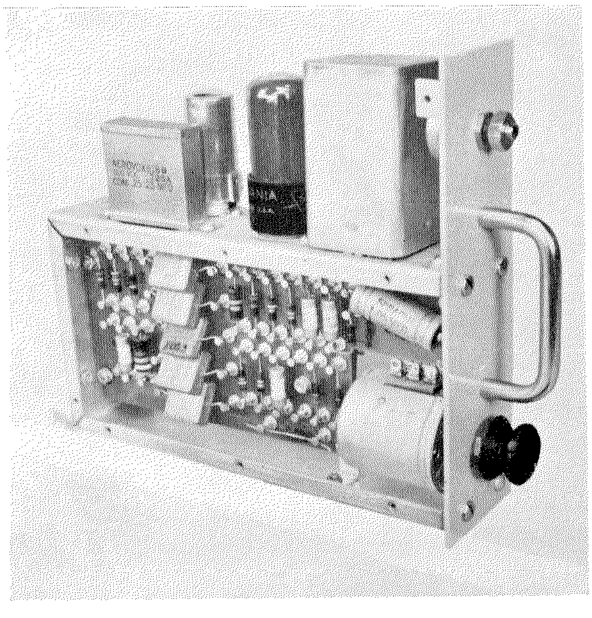
Full-size illustration of printed-circuit technique as applied to a miniature three-stage audio-frequency amplifier. Front and rear views are shown. The unit measures $1\frac{5}{16}$ by $2\frac{1}{4}$ inches. (Photo courtesy Centralab Division, Globe Union Corporation, Milwaukee, Wisconsin.)

From this situation arises a new concept of the "life" of apparatus. Most manufacturers rate the probable life of their designs on a basis of 50-percent failure, or in some cases 20-percent or 10-percent failure, averaged over a number of equipments. "Life" figures based on such concepts are of little interest to military people, since they bear no fixed relation to the factor that is all important; which might be called the "guaranteed life." This figure would represent the period during which *none* of the equipments would fail, and absolute reliability could be assumed. (Since failure is a statistical phenomenon, absolute guarantee might not be possible, but military needs might require that failures in this period be of the order of one percent or less, depending on the application.) Even though this "guaranteed life" were quite short, equipment could be serviced or replaced as required, with virtual freedom from unexpected failures.

1.4 Ease of Production

While it may appear that cost is of no importance in military apparatus, this is far from true. While money may be appropriated freely, in the final analysis neither materials nor manpower are inexhaustible and methods adapted to simple rapid production are frequently a major requirement. For example, without this approach the wholesale expenditure of complete transmitter-receiver units could not have been contemplated, and the proximity fuze would have remained a laboratory toy.

Startling progress has been made in this field, particularly in the development of methods to reduce the labor of assembling and wiring. Circuits are stamped, painted, printed, sprayed, or molded, combining in one operation that which would otherwise require many steps. In addition, these methods are well adapted to miniaturization and unitized construction. It seems certain that vigorous experimentation



Unitized modulator unit of the plug-in type of construction as used in the FTL-10-A 23-channel multiplex radio link.

along these lines will bring startling changes in both military and commercial apparatus in the next few years.

1.5 Other Trends

Several additional trends may be noted. Except where power output is needed, there is no real need for the high plate voltages customarily used. Steps to reduce operating voltages frequently simplify problems of power supply and heat dissipation.

Increasingly severe requirements in ambient conditions are being imposed by military needs. For use in the arctic, equipment should survive storage at −85 degrees centigrade for long periods and be operable at −65 degrees centigrade. In some new jet aircraft, the most desirable position for the installation of radio gear would subject it to temperatures approaching 200 degrees centigrade. Until these conditions can be met without sacrifice of performance or reliability, we cannot say we have finished our job.

Increased simplicity of operation, with automatic control over all possible points of misadjustment, will make it possible for personnel untrained in electronics to use new apparatus with ease and certainty.

In an endeavor to decrease the total number of units required in a given installation, an effort is being made to increase the frequency coverage of single transmitter and receiver units. Greater precision of channel selection will permit narrower channels, but this must be obtained without increasing excessively the number of quartz crystals in the equipment.

2. *Conclusion*

Trends in the development of electronic equipment for military use show that every designer should check his products for the application of the following principles: miniaturization, unitized construction, extreme reliability for certain "guaranteed life," and ease of production without waste of manpower or materials. When these needs have been satisfied as fully as the electronic art permits, he may be sure his design will fill the purpose for which it was intended with maximum utility.

Development of Single-Receiver Automatic Adcock Direction-Finders for Use in the Frequency Band 100-150 Megacycles per Second*

By R. F. CLEAVER

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THE PAPER describes the development of automatic direction-finders primarily intended for measuring the bearings of aircraft, on signals radiated in the frequency band 100 to 150 megacycles per second. The basic system employs fixed elevated-H Adcock antennas and a single receiver, and, with the exception of the earliest model, all instruments give cathode-ray-oscillograph indication.

An examination of the advantages and limitations of fixed-antenna automatic direction-finders employing two or more receivers leads to an account of the reasons for the adoption of a single-receiver system in the present development. The principles and evolution of this system are explained with reference to the original experimental models, whose performance is discussed.

A naval direction-finder based on one of the experimental models is described in detail, and statistics are presented to show that the probable instrumental error varies from about 0.6 to 1.25 degrees over the frequency band, after compensation for octantal error. Bearings can be measured on signals of field strength down to 7 microvolts per metre or less.

The paper concludes with a description of direction-finders for use on land. One of these, under current development, will be capable of unattended operation on two alternative frequency-channels, with bearing indication and full remote-control at points up to 20 miles distant.

• • •

1. Introduction

The development described in this paper was originally undertaken in 1941 with the primary object of increasing the speed and reliability of very-high-frequency direction-finding at aerodrome ground-stations. At that time it appeared likely that increasing traffic would shortly lead to saturation of the manual direction-finding system, which had served very well during the critical early years of the war. In the event, however, it proved impossible to develop and introduce the automatic system in time for the peak period of Royal Air Force activity, and with the completion of experimental models, the development might have been discontinued in the absence of a new incentive. This arose towards the end of 1942, in the form of a requirement of the British Admiralty for a new very-high-frequency direction-finder for use on aircraft carriers. Satisfactory tests with the second experimental instrument described later in this paper led to the adoption of a virtually identical circuit for the new naval direction-finder, now known in its production form as "Admiralty D/F Outfit FV5."

Since the end of the war it has become possible to resume the development of the system to meet specific ground-station requirements, notably unattended operation of the direction-finder with remote indication and control over many miles of land-line.

2. Choice of System

Some of the principles adopted in the developments described in this paper had not previously been applied to any great extent, and, before giving a detailed account of the work, it is necessary to examine the reasons underlying the choice of system.

For the antenna system, there was no practical alternative to the elevated-H Adcock arrangement, which is extremely convenient for wavelengths between 2 and 3 metres, and combines reasonable fredom from polarization error with adaptability for automatic indication of bearings on a cathode-ray oscillograph. The ultimate use of oscillographic indication was regarded as essential for speed of operation and for automatic indication of the presence of a signal, though it was expedient to use a different form of indication on the first model.

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* Reprinted with minor additions from *Journal of the Institution of Electrical Engineers,* v. 94, Part IIIA, n. 15, pp. 783–799; 1947.

Choice of method for automatic indication lay between the static and the continuously rotating antenna (or goniometer) systems. The former were preferred for their flexibility in speed of indication and adaptability to more satisfactory forms of bearing indication. There was also some reluctance to introduce continuously rotating machinery if this could be avoided.

Automatic direction-finders of the "static" class are in general characterized by the fact that signals related in amplitude by a tangent law are derived from two orthogonal antenna combinations, amplified with or without frequency conversion, and applied to a cathode-ray tube to produce corresponding orthogonal deflections of the electron beam. The resultant deflection can then be arranged to indicate the direction of arrival of the signal. The voltage amplification required may be up to the order of 10^8 times, and no difference in amplification of the signal components can exceed 2 per cent if the maximum quadrantal error is to be kept below 0.5 degree. Signals from the two directional antenna combinations can only give the line of the required bearing; to obtain its sense, the phases of the signals from the directional antenna combinations must be compared with that of a signal from a central omni-directional antenna. The latter signal requires corresponding amplification, without change of relative phase.

In general, the signals may be amplified in two ways; either in separate balanced amplifiers or, if previously identified by modulation, in a common amplifier with subsequent separation before application to the deflecting elements of the oscillograph. The following considerations led to the choice of the second alternative for the direction-finders described in this paper. The two systems are here distinguished as "balanced-receiver" and "single-receiver" systems.

2.1 BALANCED-RECEIVER SYSTEMS

Automatic direction-finders employing balanced receivers were first constructed by Watson Watt and Herd more than 20 years ago for research into the nature and origin of atmospherics,[1] and have since been widely used, particularly for high-frequency work. Details of some of the earlier instruments have been published[2, 3] but at the time of writing (November, 1946) information on modern equipments of this type is lacking. The following discussion is therefore based on general principles, without reference to specific developments. The relative importance of the advantages and disadvantages depends greatly on the frequency band for which the direction-finder is designed.

Advantages

A. Basic simplicity of conception.

B. Possibility of obtaining a bearing on one transmission in the presence of interfering radiation.

C. Possibility of assessing the reliability of bearing indications from the character of the oscillograph display.

Disadvantages

A. Necessity for a high degree of balance between the receivers in respect of gain, phase shift and bandwidth. This involves the use of special receivers with provision for stage-by-stage equalization between them. Manual gain-controls must be ganged and provided with matching adjustments. Automatic gain-control necessarily involves considerable difficulty and is not used in instruments that have been described in the literature.

B. Necessity for frequent checking of the equalizing adjustments, whose unavoidable variations lead to quadrantal error and ellipticity of indication.

C. Limitation in the improvement in signal-to-noise ratio which can be gained by the use of selective circuits, governed by considerations of frequency stability and consequent difficulty in maintaining the necessary phase and amplitude balance.

D. Directional accuracy is impaired by any inequality between transmission lines connecting the antennas and receivers, and by any stray pick-up on them. Difficulties in these respects increase with frequency.

E. The form of bearing display obtained is not suitable for remote presentation, nor is the equipment itself suitable for unattended operation.

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[1] R. A. W. Watt, and J. F. Herd, "An Instantaneous Direct-reading Radio Goniometer," *Journal of the Institution of Electrical Engineers*, v. 64, p. 611; 1926.

[2] R. A. W. Watt, J. F. Herd, and L. H. Bainbridge-Bell, "Applications of the Cathode-Ray Oscillograph in Radio Research," 1st Edition, His Majesty's Stationery Office, London, 1933; pp. 123–242.

[3] Staff of the Radio Research Station, "A Short-Wave Cathode-Ray Direction Finding Receiver," *Wireless Engineer*, v. 15, p. 432; 1938.

The first three disadvantages are fundamental and peculiar to balanced-receiver systems, while the other two may apply also to single-receiver systems, to an extent depending on their design.

2.2 Single-Receiver Systems

All the direction-finders described in this paper are based on the single-receiver principle, and are thus free from the fundamental disadvantages of the earlier system. The fact that they lack the special qualities listed as advantages of the balanced-receiver system is insignificant for the purposes of the present development, as will be clear from the following considerations.

First, the theoretical simplicity of the balanced-receiver principle is largely illusory owing to the formidable number of adjustments which are necessary to make the system work, and the frequency with which they must be checked. On the other hand, if one accepts the initial complication of modulation and demodulation involved in the single-receiver system, the circuits required can be made very stable and do not require continuous adjustment. A normal communication receiver with automatic gain-control may be employed in the common channel.

Secondly, the inability of the single-receiver system to indicate correct bearings in the presence of an interfering signal is harmless so long as the interference is discontinuous and of short duration, and so long as its presence can be detected. These conditions are invariably satisfied in the very-high-frequency aeronautical band. Significant interfering signals are revealed either by their modulation, by an audible heterodyne note, by an abrupt change in bearing indication during the observation, by the presence of a bearing indication preceding or following the wanted transmission, or by any combination of these effects. Under bad conditions, a bearing indicator with a high speed of response gives the required information in short intervals between interfering signals.

Finally, the fact that a single-receiver instrument under automatic gain-control may not give an indication of propagation conditions likely to lead to serious error is only of importance under the sort of conditions liable to be encountered at frequencies below, say, 50 megacycles per second, and it is of little consequence in direction-finding on very-high-frequency signals from aircraft. At all material times these signals contain a substantial vertically polarized component and have a large angle of incidence. In general, conditions capable of giving bad polarization errors are only to be expected when the aircraft has a high angle of elevation, and when this is the case its speed and proximity will generally preclude accurate direction-finding in any case.

2.2.1 Identification of Signal Components in Single-Receiver Systems

Various modulation schemes have been suggested for signal-component identification. In the 3-tone system described by C. F. A. Wagstaffe,[4] the signals from the two directional elements and the omni-directional element of the antenna system are modulated at different audio-frequencies. After amplification and detection, the 3 audio-frequency components are separated by filtering, and those corresponding to the directional elements have their frequencies changed to that of the third. The 3 components of common frequency are used to produce a bearing display similar in some respects to that of a balanced-receiver system. Although free from the fundamental disadvantages of the latter, the 3-tone system as described has certain shortcomings and, in particular, the use of 3 modulation-frequencies is a complication which can be avoided.

All the direction-finders described in this paper make use of the 2-tone system due to C. W. Earp,[5] in which balanced modulation at different frequencies is applied to the signals from the 2 directional-antenna elements only. This results in suppression of the carrier which is later restored in constant phase by the addition of a signal derived from the omni-directional antenna, the combined signal then passing to the common receiver. The use of balanced modulators renders the carrier level substantially independent of azimuth, and so permits a greater average depth of modulation. The output from the receiver includes 2 audio-frequency components, whose relative amplitudes depend on the line of bearing of the received signal and whose line phases, relative to those of the original modulating tones, depend on its sense.

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[4] British Patent No. 496239.

[5] British Patent No. 490940.

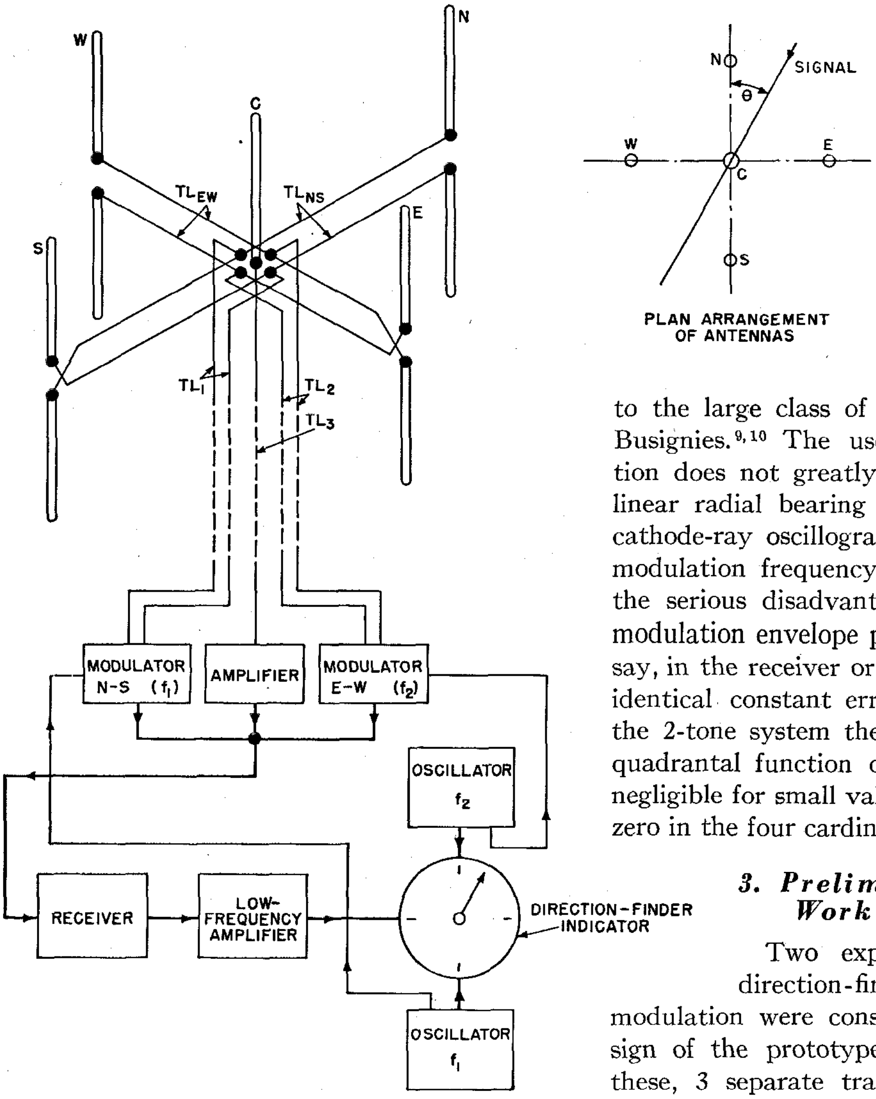

Fig. 1—First experimental model—block circuit diagram.

Two methods of indicating the bearing are described in the original patent, one employing a cathode-ray oscillograph and the other an electro-dynamometer type of instrument. More recently, an improved form of oscillograph display has been developed and is now used exclusively on the direction-finders here described.

A variant of the 2-tone system uses quadrature components of a single tone for balanced modulation of the directional signals. The phase of the receiver output at modulation frequency is then a linear function of the direction of arrival of the signal, which can therefore be determined by phase comparison with the modulating tone. The process is electrically analogous to continuous rotation of a loop antenna, dipole pair or goniometer search-coil, and thus yields an instrument related to the automatic direction-finders[6] proposed by Braillard and Goldschmidt,[7] Marique,[8] and others, and to the large class of instruments developed by Busignies.[9,10] The use of quadrature modulation does not greatly simplify the circuit if a linear radial bearing display is required on a cathode-ray oscillograph, and moreover, if the modulation frequency is high, the method has the serious disadvantage that a change in the modulation envelope phase-shift by an angle α, say, in the receiver or display circuits causes an identical constant error in all bearings. With the 2-tone system the corresponding error is a quadrantal function of $(1-\cos\alpha)$; it is quite negligible for small values of α and is in any case zero in the four cardinal directions.

3. Preliminary Experimental Work

Two experimental single-receiver direction-finders employing 2-tone modulation were constructed prior to the design of the prototype model. In the first of these, 3 separate transmission lines were employed between the antenna system and the modulators and amplifier in the direction-finder office, and the band-width which could be covered without antenna and line adjustments was very restricted. In the second model, this difficulty was overcome by transferring the modulators and amplifier to the masthead. At the same time, the method of indicating the bearing was improved by the substitution of a cathode-ray oscillograph for the original electromechanical indicator. The second model was used for much of the experimental field-work and many of the results quoted in Section 4.5 were obtained with it.

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[6] R. Keen, "Wireless Direction Finding," 3rd Edition, Iliffe, London, 1938; pp. 688–700.

[7] French Patent No. 516295.

[8] J. Marique, "Un nouveau radiogoniomètre à lecture directe," *L'Onde Electrique*, v. 10, p. 355; 1931.

[9] H. G. Busignies, "Un nouveau radio-compas," *L'Onde Electrique*, v. 9, p. 397; 1930.

[10] H. G. Busignies, "The Automatic Radio Compass and its Application to Aerial Navigation," *Electrical Communication*, v. 15, pp. 157–172; October, 1936.

3.1 First Experimental Model

This was designed to operate in the frequency band 100 to 125 megacycles per second, using a British Air Ministry receiver, Type R1132A.

3.1.1 Theory

This will be considered in some detail since the basic principles were preserved in all subsequent equipment, though the circuit technique was changed radically.

The form and inter-connection of the fixed Adcock antenna-array were conventional and will be apparent from the block circuit diagram and plan of the system, Fig. 1. The diagonal transmission lines TL_{NS} and TL_{EW} were made equal to a half-wavelength in the mid-band frequency (112 megacycles per second), and transmission lines TL_1, TL_2 and TL_3 were equalized electrically at the frequency of alignment. The modulators and amplifier were located in the direction-finder office and had a common output-circuit connected to the receiver.

When the antenna system is exposed to radiation of wavelength λ having a vertical electric field strength $\mathcal{E}_V$ and an azimuthal direction of arrival θ, it can be shown by the aid of well-known theory that the voltages applied to the amplifier and the N–S and E–W modulators are respectively of the form

$$e_C = k_C \mathcal{E}_V \sin(\omega t - \alpha_3) \tag{1}$$

$$e_{NS} = k_{NS} \mathcal{E}_V \sin(2\pi d \cos\theta/\lambda) \times \sin(\omega t + \tfrac{1}{2}\pi - \beta_{NS} - \alpha_1) \tag{2}$$

$$e_{EW} = k_{EW} \mathcal{E}_V \sin(2\pi d \sin\theta/\lambda) \times \sin(\omega t + \tfrac{1}{2}\pi - \beta_{EW} - \alpha_2), \tag{3}$$

where k_C, k_{NS} and k_{EW} are constants of the system, α_1, α_2, α_3, β_{NS} and β_{EW} are the phase shifts in transmission lines of corresponding suffix and $2d$ is the diagonal of the antenna system.

If

$$\alpha_1 = \alpha_2 = \alpha_3 = \alpha, \quad \beta_{NS} = \beta_{EW} = \tfrac{1}{2}\pi$$

and

$$k_{NS} 2\pi d/\lambda = k_{EW} 2\pi d/\lambda = k,$$

and if

$$2\pi d/\lambda \not> 1 \text{ (say)},$$

the expressions for the input voltages reduce to

$$e_C = k_C \mathcal{E}_V \sin(\omega t - \alpha) \tag{4}$$

$$e_{NS} \simeq k \mathcal{E}_V \cos\theta \sin(\omega t - \alpha) \tag{5}$$

$$e_{EW} \simeq k \mathcal{E}_V \sin\theta \sin(\omega t - \alpha). \tag{6}$$

The sign of approximation in (5) and (6) indicates that octantal spacing-error is being ignored; for exact equality, $2\pi d/\lambda$ would have to be indefinitely small.

The signals (5) and (6) are modulated at frequencies f_1 and f_2 with suppression of carrier, and the signal (4) is amplified, all the outputs being combined in the common circuit to yield a signal of the form

$$e \simeq k_C' \mathcal{E}_V \sin(\omega t - \alpha)[1 + (k'/k_C') \times (\cos\theta \sin 2\pi f_1 t + \sin\theta \sin 2\pi f_2 t)], \tag{7}$$

where k_C' and k' are new constants including the gains of the amplifier and modulators respectively, those of the latter being assumed to be equal. Phase shifts in the modulators and amplifier are assumed equal and omitted from (7) to avoid complication.

Expression (7) represents the modulated signal applied to the receiver. The relative gains of the modulators and amplifier are assumed to be regulated to ensure that the maximum instantaneous modulation depth $k'\sqrt{2}/k_C'$ (which is obtained for bearing angles making $\cos\theta = \pm\sin\theta$) is less than the maximum modulation depth that can be handled by the receiver. In a typical case, allowing for a factor of safety, k'/k_C would not be allowed to exceed 0.5.

The output from the receiver contains components of current at the modulation frequencies f_1 and f_2 which are of the form

$$i \simeq K f(\mathcal{E}_V)[\cos\theta \sin(2\pi f_1 t - \psi_1) + \sin\theta \sin(2\pi f_2 t - \psi_2)], \tag{8}$$

where K is a new constant including the gain of the receiver and the load impedance, $f(\mathcal{E}_V)$ depends on the law of the detector and ψ_1, ψ_2 represent the modulation-frequency envelope phase-shifts through the receiver, and phase-shifts in the low-frequency circuits.

The current i passes through the signal coil of the bearing indicator whose field coils N–S and E–W are supplied with constant current i_1 and i_2, derived from the modulating oscillators, and adjusted to be in phase with the corresponding components of the signal current. The instrument has a freely rotatable element carrying a pointer indicating on a 360-degree scale. If ϕ is the pointer reading, the law of the instrument is such that synphased currents in the signal coil and the N–S field-coil exert a torque on the moving element approximately proportional to their product and to $\sin \phi$. Similarly, synphased currents in the signal coil and the E–W field-coil exert a corresponding torque proportional to $-\cos \phi$.

Influenced by currents i, i_1 and i_2, the moving element is acted upon by two torques varying at the modulation frequencies and takes up a mean angular position ϕ in which the average value of the sum of the torques is zero. This position is defined by

$$\text{av. value of } (i \cdot i_1 \sin \phi) - \text{av. value of } (i \cdot i_2 \cos \phi) = 0. \quad (9)$$

Thus for the mean position

$$\cos \theta \sin \phi - \sin \theta \cos \phi \simeq 0 \quad (10)$$

or

$$\phi \simeq 0.$$

The mean position of the pointer therefore indicates the bearing of the incoming signal, and as the rotatable element is in general acted upon by torques varying at different frequencies, it vibrates at a frequency equal to their difference with an amplitude which depends on the strength of signal. This vibration is limited to the order of ±1 degree or less by inertia, friction and eddy-current damping.

Components of the receiver output-signal at frequencies other than f_1 and f_2, including noise voltages, can only produce oscillatory or transient movements of the indicator. These movements are small unless the signal components at f_1 and f_2 are very weak, corresponding with conditions close to maximum range.

3.1.2 Practical Details

The four dipoles were resonant at 112 megacycles per second and their diagonal spacing, 76 centimetres, caused a maximum octantal error of 2 degrees at the same frequency. The antenna structure was carried on a 25-foot steel tubular mast, rotatable through 360 degrees to allow investigation of instrumental error on a single incoming signal.

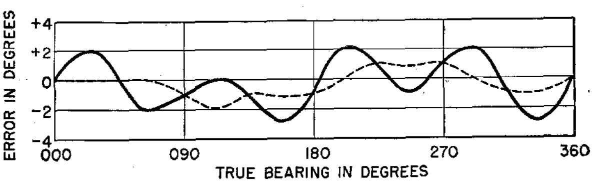

Fig. 2—First experimental model—instrumental error. Full line—uncorrected. Broken line—corrected for 2-degree (maximum) octantal error.

Transmission lines TL_{NS} and TL_{EW} were made from twin circular mineral-insulated copper-clad cable having a characteristic impedance of about 70 ohms and TL_1, TL_2 were each made from two lengths of the same material parallel at both ends. TL_3 was a coaxial cable of similar construction. Interchangeable link sections at the receiving end were used to adjust the electrical lengths of these lines.

The lines were coupled inductively to the amplifier and modulators, which comprised respectively a single CV1091 (EF50) high-frequency pentode, and two pairs of valves of the same type in push-pull input, parallel output connection. The five valves worked into a common output circuit connected over a short line to the receiver. Modulating voltages at 45 cycles per second (f_1) and 60 cycles per second (f_2) were applied to the suppressor grids, and differential cathode bias-control on the modulators provided overall compensation for quadrantal error.

3.1.3 Performance

The instrumental accuracy was measured on a local radiated signal by observing the "tracking" accuracy of the bearing indicator as the antenna system was rotated through 360 degrees in small steps, using a signal radiated by a field oscillator some 50 to 100 feet from the direction-finder antennas. This method was also used on subsequent models. A typical instrumental error curve is shown in Fig. 2.

With physically equal antenna lengths and optimum adjustments to lines and modulators, the maximum error at the frequency of alignment was nearly 2 degrees after compensation for octantal error. Quadrantal error increased to some 2 degrees (maximum) for frequency changes of ±3 megacycles per second, giving a maximum error of about 4 degrees at the least favourable frequency and azimuth in any sub-band about 6 megacycles per second wide between 100 and 125 megacycles per second.

The sensitivity was measured in terms of the minimum field strength required to give a first-class bearing indication, which was somewhat arbitrarily defined as one in which random noise caused a pointer fluctuation of ±5 degrees or less. At mid-band, the field strength required was about 6 microvolts per metre, with the modulator gains adjusted to give an instantaneous maximum modulation depth of 60 per cent.

3.1.4 Limitations

A. *Frequency Coverage.*—For frequencies differing from that of alignment by more than 3 megacycles per second, mis-phasing between sideband and carrier components of the signal caused increasing quadrantal error, loss of sensitivity and in extreme cases, sense reversal.

B. *Presentation of Bearings.*—Friction and mechanical unbalance affected the accuracy of bearing indication on weak signals, and inertia and damping limited the speed of response. There was no effective indication of the strength or absence of a signal.

C. *Speech Reception.*—Low modulation-frequencies were dictated by the design of the bearing indicator and resulted in poor intelligibility of speech reproduction during direction-finder operation.

3.2 SECOND EXPERIMENTAL MODEL

The above limitations were largely overcome in a second model whose circuit was later adopted with little alteration for the Admiralty D/F Outfit FV5 described below. The block schematic diagram (Fig. 3) applies equally to both.

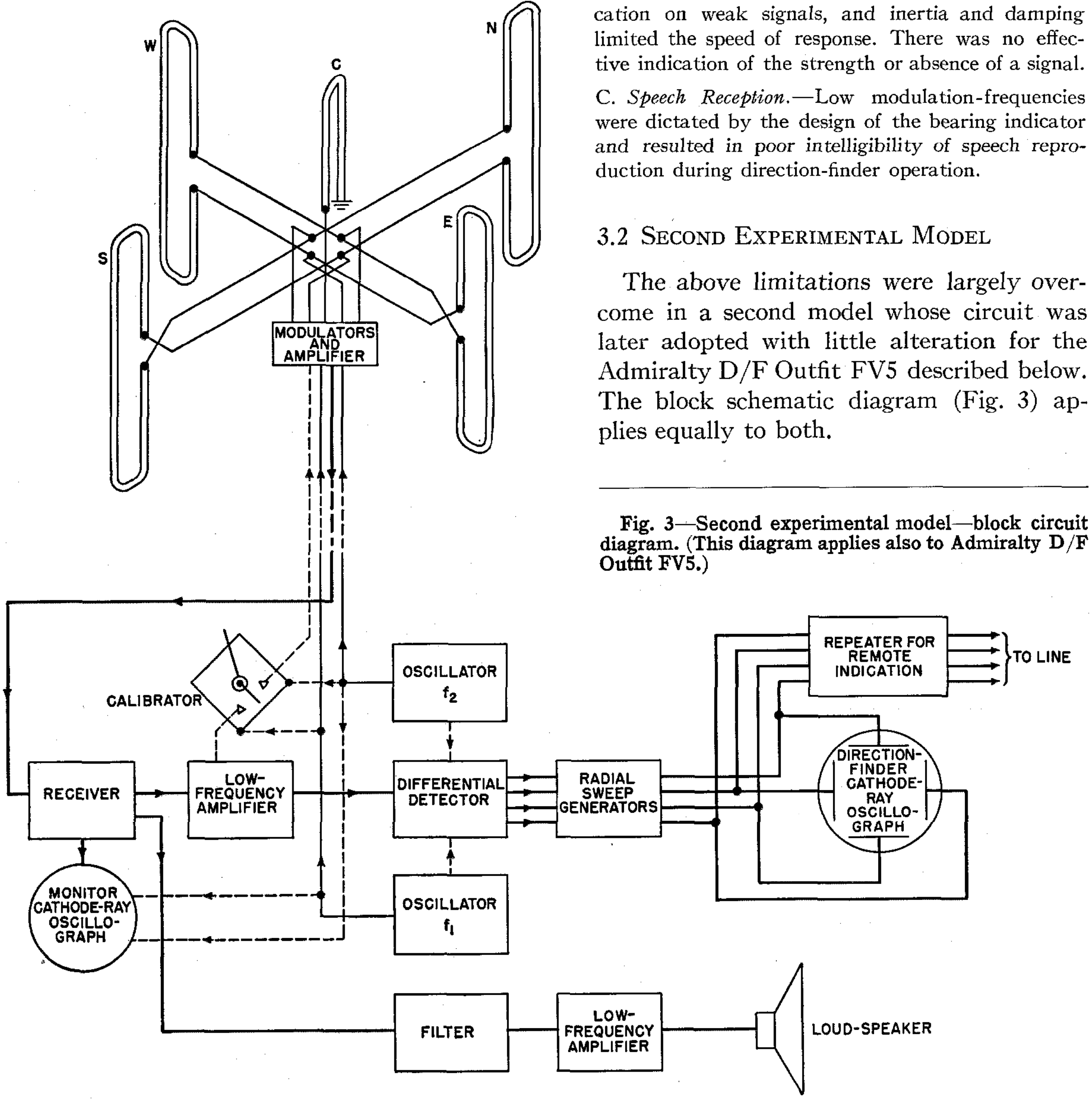

Fig. 3—Second experimental model—block circuit diagram. (This diagram applies also to Admiralty D/F Outfit FV5.)

3.2.1 Improvements in Technique

A. *Frequency Coverage.*—Folded antennas were adopted with direct connection to the amplifier and modulators, which were transferred to the mast-head. This change was made possible by the use of CV53 (Standard Telephones and Cables type S26A) grounded-grid triodes working into a band-pass output circuit, connected over a single coaxial line to the receiver below. A useful increase in accuracy resulted from the improved shielding given by the grounded-grid arrangement.

B. *Presentation of Bearings.*—To obtain cathode-ray-oscillograph indication, the modulation-frequency components of receiver output [expression (8), above] had first to be converted to a common frequency in separate channels, and then filtered to remove unwanted signal components and improve signal-to-noise ratio. The common frequency finally chosen was zero, the conversion being effected in differential detectors working into resistance-capacitance filters of adjustable time-constant. Although this arrangement is exceedingly convenient from the point of view of the designer, bearing indication by the steady displacement of the oscillograph beam was found to be psychologically unsatisfactory; some form of radial sweep appeared to be necessary. Some improvement was obtained at first by modulating the deflection sensitivity of the oscillograph, but this arrangement was later abandoned in favour of an electronic switching circuit[11] whereby the "spot" indication was converted to a full radial trace.

C. *Speech Reception.*—If the modulation frequencies and the difference between them were made high enough to eliminate all audible interference, large and very unequal modulation-envelope phase-shifts would occur in the receiver. This would have been undesirable for high stability of phase adjustment, and a satisfactory compromise was found in the use of 5 kilocycles per second for f_1 and 6 kilocycles per second for f_2, with a sharp band-elimination filter in the speech circuit, centred on the difference frequency, 1 kilocycle per second, and a low-pass filter with cut-off at about 3 kilocycles per second to eliminate modulation-frequency components. The filters scarcely affected the intelligibility of speech.

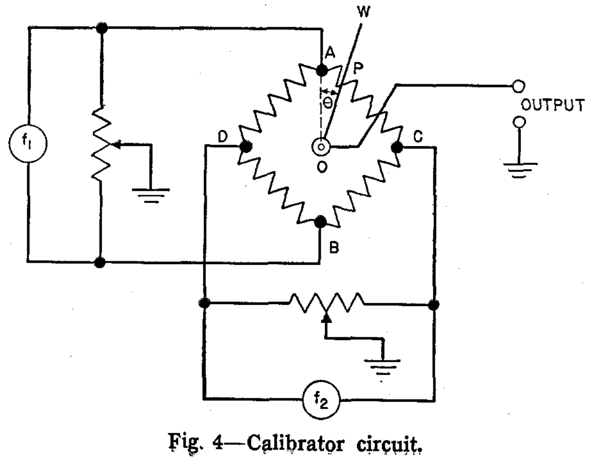

Fig. 4—Calibrator circuit.

3.2.2 Monitoring and Testing

As in the first model, the antennas were mounted on a mast, which could be rotated through 360 degrees for the measurement of instrumental error. A further check on the behaviour of the antenna system was obtained by the inclusion of a small monitor oscillograph (CV967) to indicate the modulation depth and permit estimation of the relative phase angles between carrier and sideband-resultants.

For checking the display circuits, a device was incorporated to simulate the modulation-frequency components of receiver output corresponding to any bearing θ. It utilized a simple geometrical property of the square which does not appear to be generally known and has been found to have a number of other useful applications.

The arrangement is illustrated in Fig. 4. Equal balanced-to-ground voltages at frequencies f_1 and f_2 are applied to the diagonally opposite points A, B, and C, D of a closed network of four equal, straight, uniform resistance elements arranged physically in the form of a square and resembling a Wheatstone bridge in which the battery and galvanometer are replaced by the sources of voltage at the two modulation-frequencies. A straight wiping contact arm OW is pivoted at O, the centre of the square, and by rotation in its plane can make contact at any point P on the network. If angle AOP, the angular position of the arm, is denoted by θ it can readily be shown that the point of contact P divides the element AC of the network so that $AP/PC = \tan\theta$. Thus the voltage picked up by the contact arm contains components at frequencies f_2 and f_1 in the ratio ($\tan\theta$), and is therefore equivalent to the receiver output corresponding with a signal from a direction θ. Symmetrical deformation of the resistance network can be made to simulate various types of periodic error. For example, a circular network gives an octantal error of maximum value 4 degrees, of sign opposite to that of the octantal spacing error in an Adcock antenna system.

It may be noted that the device provides a facility similar to that of the angular calibrator of Watson Watt, Herd and Bainbridge-Bell,[12] though in the present equipment it is not used to calibrate the oscillograph. Although generally referred to as a "calibrator," it was usually employed for preliminary adjustments and functional tests on the display circuits. By switching off the antenna modulators and using the calibrator output to modulate any carrier wave at the grid of the central antenna amplifier the functional test could be extended to cover the receiver.

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[11] British Patent No. 590260.

[12] Reference 2, p. 162.

3.2.3 *Mechanical Construction*

The direction-finder office equipment was rack-mounted and with the exception of the cathode-ray-oscillograph units was accommodated on one 19-inch rack, 6 feet high. In later developments, the need for front accessibility and exceptionally rugged construction led to a considerable increase in size.

3.2.4 *Performance*

In performance, the second instrument resembled production models described in detail later. The improvement over the first model is indicated by an increase in working bandwidth by a factor of about four, the band 100 to 125 megacycles per second being covered without adjustment to antennas or lines. Later tests showed that the band could be increased to 100 to 150 megacycles per second before serious difficulties with sideband phasing were encountered. For a much greater extension it is probable that folded dipoles would have to be abandoned.

For practical purposes, the sensitivities of the two experimental models were similar, though the use of different methods of bearing indication prevented direct quantitative comparison.

4. *British Admiralty D/F Outfit FV5*

This direction-finder was developed for use on aircraft carriers and covered the frequency band 100 to 150 megacycles per second. The circuit is essentially the same as that of the second experimental model (Fig. 3) except that a single output circuit with remote tuning-control was used in the antenna system. A prototype known as "Admiralty D/F Outfit FV5X" was made in small numbers and covered the band 115 to 145 megacycles per second with a band-pass output circuit. In other respects it resembled the production equipment.

4.1 ANTENNA SYSTEM

The construction of the antenna system is shown in the photographs, Figs. 5 and 6, and the circuit in schematic form in Fig. 7.

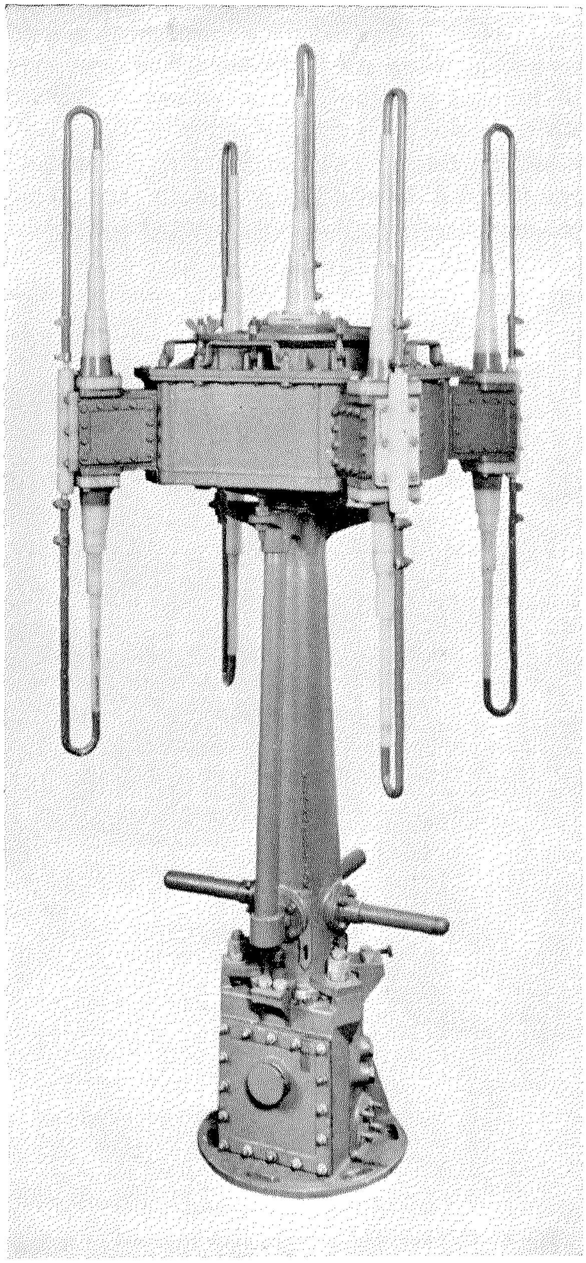


Fig. 5—FV5 antenna system.

4.1.1 Mechanical Construction

Polythene-covered folded-dipole antennas are mounted on short stubs projecting from the corners of a cast aluminium-alloy box whose hinged lid carries the central folded monopole. The box is supported by a short tubular column mounted on a junction box and capable of limited rotation for test purposes. The whole assembly is water-tight and is normally bolted to the top of a steel lattice mast, hinged at its base, which is carried on a sponson at a level somewhat below the flight-deck. The mast may be lowered outboard into a horizontal position when the direction-finder is not in use.

Within the box, the air-dielectric "diagonal" transmission lines actually run in cast metal shields of rectangular section fixed along the inner sides. Plug-in modulators and central amplifier facilitate valve changing; their arrangement can be seen in Fig. 6. The box also contains the common output circuit and a mains transformer for valve heaters. The total weight of the water-tight antenna assembly as illustrated in Fig. 5 is 160 pounds.

Diagonal spacing of antennas is 76 centimetres, giving maximum octantal errors of 1.6 degrees at 100 megacyles per second and 3.8 degrees at 150 megacycles per second, normally included in the site error calibration. When the direction-finder is used on land it is advantageous to use a cathode-ray-oscillograph scale compensated for octantal error. Such compensation is appropriate on instruments designed for direct-ray operation, when great accuracy is not required on high-angle signals.

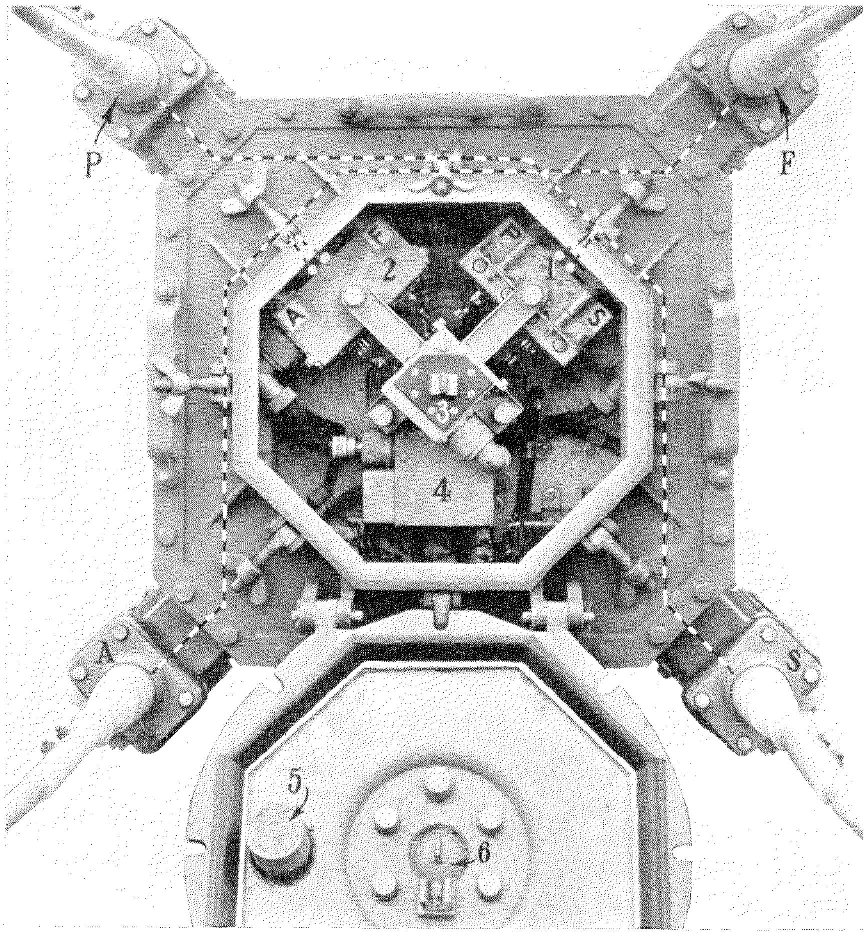


Fig. 6—FV5 antenna system—view from above with lid open. (1) Port-to-starboard modulator. (2) Fore-to-aft modulator. (3) Central amplifier. (4) Common output circuit. (5) Silica-gel desiccator. (6) Central antenna contact.

4.1.2 Circuit

The conventional Adcock arrangement was retained (Fig. 7) but the use of folded antennas permitted direct connection to modulators and amplifier and so assisted in the maintenance of good sideband phasing and quadrantal balance over the frequency band. Within the band the folded antennas show a rather smaller variation in phase angle than plain antennas, and are therefore proportionately less critical as to length.

As in earlier models the diagonal transmission lines are designed to match the antennas (300 ohms) and to have a total electrical length of about half a wavelength at the mid-band frequency.

Resistive shunts are used to give the balanced modulators an input impedance of about 150 ohms at mid-band, the valve input capacitances being annulled at this frequency by shunt inductances. Centre-taps on the latter are by-passed to ground to reduce "antenna effect" and connected to a differential cathode bias control in the direction-finder office for quadrantal balance adjustment. Modulation is applied at the control grids which are effectively grounded for signal-frequency currents by condensers integral with the valve housings.

The central antenna and its inverted amplifier are approximately matched by virtue of negative feedback so that no resistive loading is necessary in this case. The amplifier input capacitance is annulled at mid-band as in the case of the modulators, and provision is made for applying modulating voltage to the grid from the calibrator for test purposes (Section 3.2.2).

The output circuit, tuned by remote control from the direction-finder office, is common to all five valves and at the optimum tapping a single coaxial line to the receiver is connected. Any slight stray signal picked up on this line merely changes the amplitude, or shifts the phase of the carrier component to a small extent, and therefore has little or no effect on the accuracy of direction-finding.

4.1.3 Factors Affecting Accuracy and Sensitivity

The accuracy of the antenna system depends on a number of factors of which the following are the most important:

A. *Mechanical Construction.*—The general use of machined castings for the structural work ensures adequate geometrical accuracy. With air-dielectric diagonal transmission lines the maximum reciprocal error which can be produced by (electrical) eccentric connection of the balanced modulators is x/d radians, where x and $2d$ are the eccentricity and antenna diagonal respectively, in the same units. For 0.5-degree maximum error, this requires that the eccentricity should be less than about 3 millimetres; the general accuracy of construction is such that adjustment of the point of connection during factory tests is not necessary.

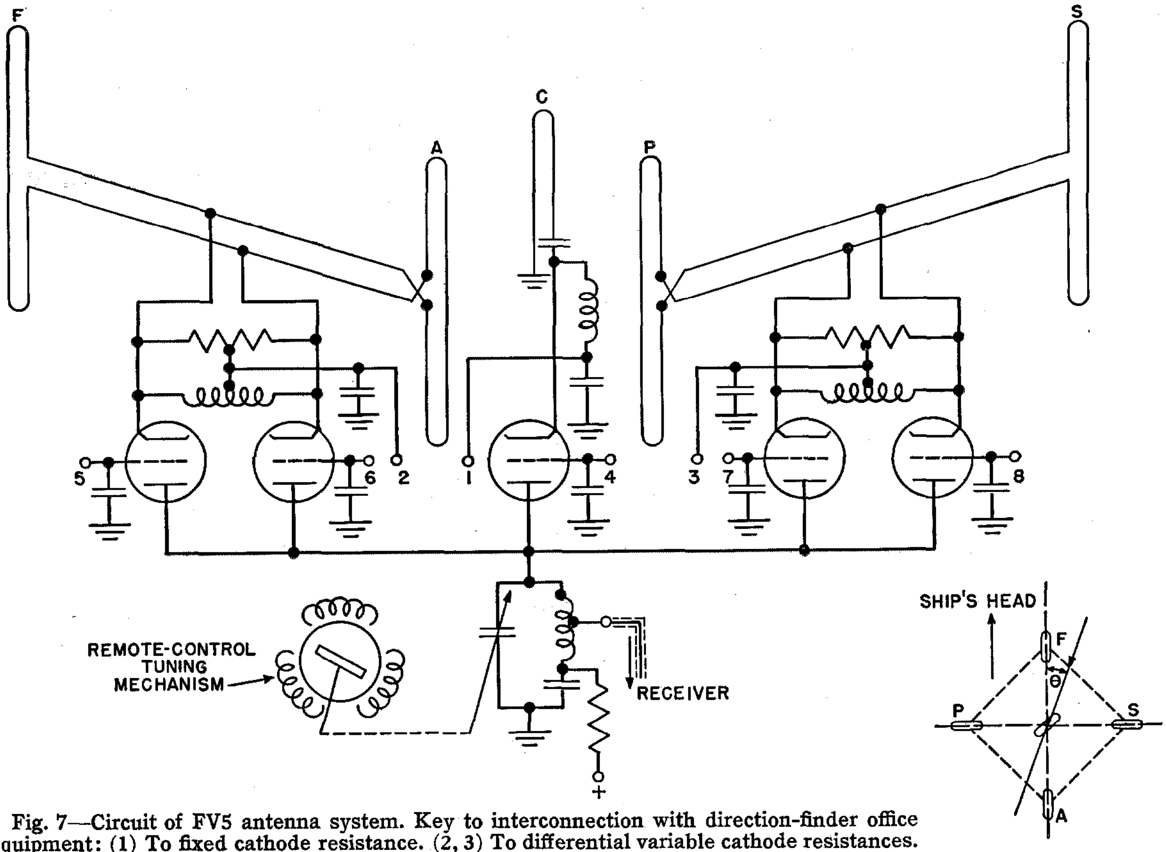


Fig. 7—Circuit of FV5 antenna system. Key to interconnection with direction-finder office equipment: (1) To fixed cathode resistance. (2, 3) To differential variable cathode resistances. (4) From ground (normal operation), or from calibrator (modulation test). (5, 6) From 5 kilocycle-per-second oscillator. (7, 8) From 6 kilocycle-per-second oscillator.

B. *Sideband-to-Carrier Phasing.*—This is kept as close as practicable over the frequency band in order to maintain sensitivity and minimize the effect on accuracy of unwanted modulation components such as might result, for example, from any residual carrier feedback into the modulators.

C. *Quadrantal Balance of Modulators.*—This is the only factor affecting accuracy that is under control in a working direction-finder of this type. It is adjusted by a method described in detail in Section 4.4.

D. *Modulation Depth.*—For good sensitivity and signal-to-noise ratio, and freedom from stray effects, the modulation depth should be as high as possible consistent with the avoidance of octantal error. In practice, a factor of safety must be allowed for variations in replacement valve characteristics.

4.2 DIRECTION-FINDER OFFICE EQUIPMENT

This is mounted in a rack combined with the operator's desk and is illustrated in Fig. 8. The receiver, control panel, bearing indicator and monitor oscillographs are carried on the sloping panel immediately behind the desk, the remaining low-frequency circuits being mounted on hinged sections of the rack above. Power supply units are mounted below the desk.

4.2.1 Common Amplification Channel

This part of the circuit is conventional and comprises a superheterodyne receiver (Admiralty Type P47) having an intermediate-frequency band-width of 90 kilocycles per second (to 6 decibels down) followed by a low-frequency amplifier which drives the differential detectors.

The receiver has two stages of radio-frequency amplification and three stages of amplification at an intermediate frequency of 9.72 megacycles per second. The supersonic heterodyne is derived from a crystal oscillator with frequency multiplication by a factor of 18. For direction-finder operation, the receiver uses biased amplified automatic gain-control. At mid-band, the output signal-to-noise ratio is 20 decibels for an input signal of 3 microvolts modulated 30 per cent at 1000 cycles per second, in series with 100 ohms. The envelope phase-shift through the receiver at modulation frequencies (chiefly in the intermediate-frequency amplifier) is about 25 degrees.

The low-frequency amplifier includes two high-gain pentode stages (CV1935) under biased automatic gain-control developed from the output of the final push-pull stage (CV1052). The first stage is carefully shielded because stray pick-up at modulation frequencies cannot be

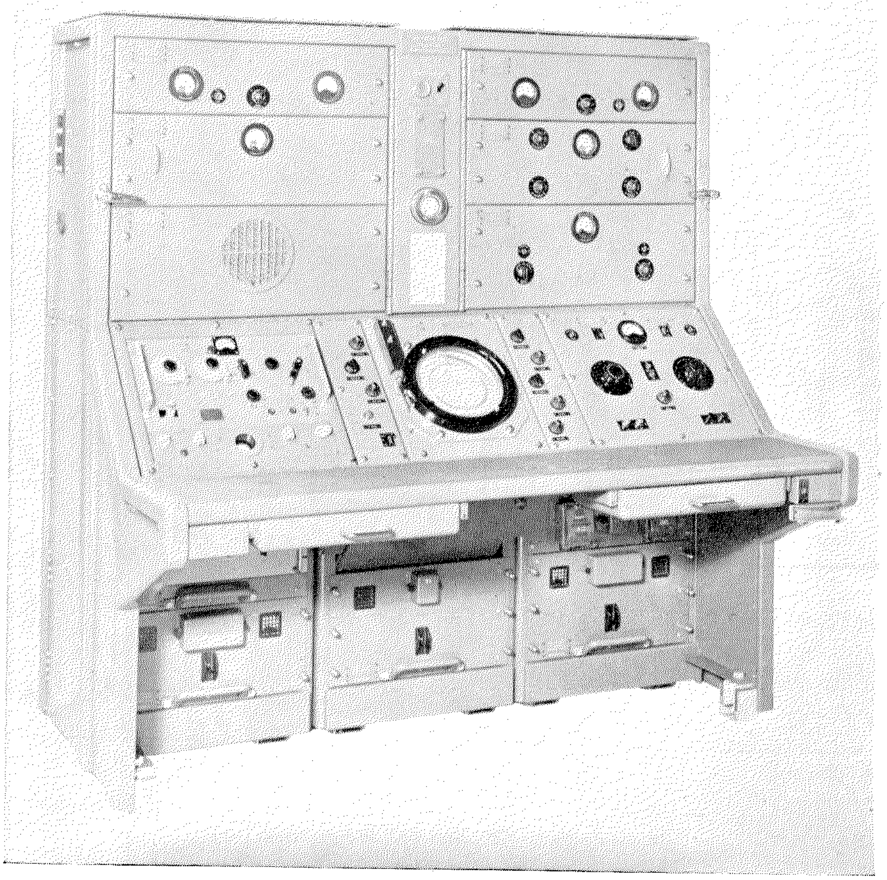


Fig. 8—Admiralty D/F Outfit FV5—main office equipment.

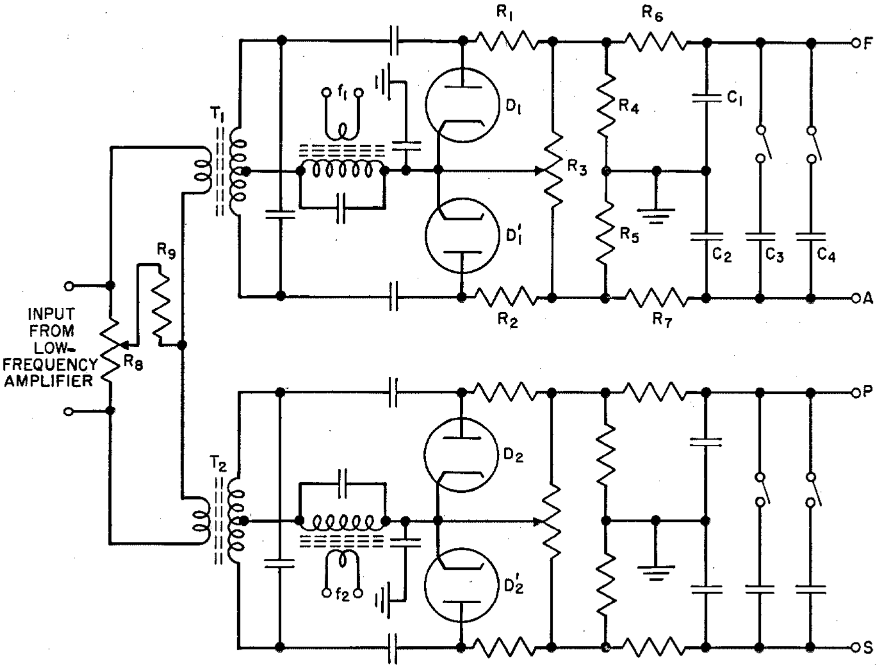


Fig. 9—Differential detector circuit.

balanced out at a later stage when automatic gain-control is used.

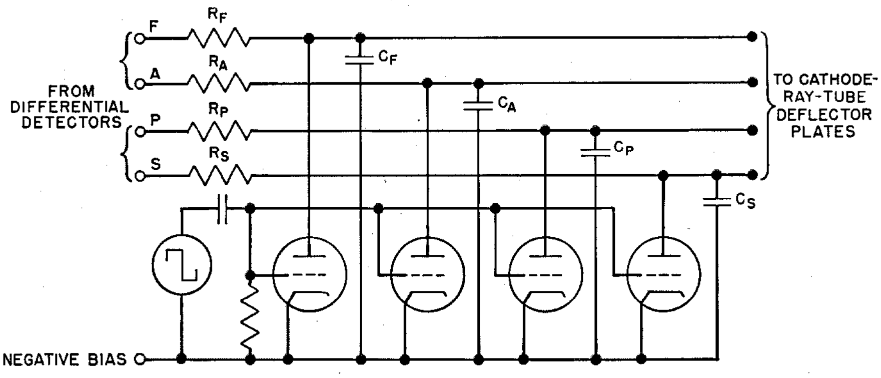
4.2.2 Separation of the Orthogonal Deflection Components of the Signal

The low-frequency output from the common channel consists mainly of components at the modulation frequencies f_1 and f_2 (5 and 6 kilocycles per second). These are separated and converted to zero frequency in the differential detectors, illustrated in principle in Fig. 9. Input transformers T_1, T_2, tuned to f_1 and f_2 apply voltages which may be up to 200 volts (root mean square) to pairs of diode peak detectors, D_1D_1' and D_2D_2' respectively. Each pair of diodes is also supplied in parallel with a "reference voltage" or zero-beat heterodyne signal, derived from the corresponding modulating oscillator. The reference voltages are initially adjusted to be in line phase with the signal component of corresponding frequency. The direct-current output from each detector can be regarded as a zero-frequency voltage proportional to that component of the signal from which it is derived, and having a "phase" (polarity) which is determined by the phase of that component, in the sense that a reversal of one causes a reversal of the other. To ensure amplitude proportionality, the magnitude of each reference voltage must be at least half that of the corresponding push-pull input signal. Referring to the f_1 detector, Fig. 9, the potential divider R_4–R_5 determines the mean output potential with respect to ground while the adjustable potential divider $R_1R_3R_2$ is a cathode-ray oscillograph centering control which equalizes the F and A output potentials in the absence of signal. Each reference-voltage phase is adjusted to give maximum cathode-ray oscillograph deflection in the appropriate direction, and the adjustment is stable within 1 degree over long periods. Misphasing causes only quadrantal error on weak signals, but an octantal component appears when the input signal becomes comparable in amplitude with the reference voltage.

Initial quadrantal error due to unbalance between differential detectors and unequal deflection sensitivities in the cathode-ray oscillograph is compensated by the differential shunt resistance network R_8–R_9.

So far, we have only considered the action of the detector on input signals of the desired frequency. For all others, the differential voltages between the diode anodes alternate at the corresponding difference-frequency, and unless this is extremely low the effect on the output voltage is completely eliminated by the resistance-capacitance filter combination $R_6C_1C_2R_7$. This has a time-constant of about 0.03 second and allows adequate response to transmissions as short as 0.1 second. The speed of indication so obtained is high enough for all telephony-modulated transmissions and is generally used under conditions of interference. A shorter time-constant could be provided, if required for keyed transmissions.

Fig. 10—Convertor circuit.


F
A
P
S
R_F
R_A
R_P
R_S
C_F
C_A
C_P
C_S
FROM DIFFERENTIAL DETECTORS
TO CATHODE-RAY-TUBE DEFLECTOR PLATES
NEGATIVE BIAS

For most purposes a time-constant of about 0.3 second is preferred and is obtained by addition of the condenser C_3. A time-constant of about 1 second is sometimes advantageous on extremely weak signals and is obtained by switching in condenser C_4. The "slow" indication so obtained is not suitable for use when interference is likely.

The three speeds of indication—"fast," "medium" and "slow"—correspond to the use of filters of bandwidth (to 3 decibels down) 10, 1 and 0.3 cycles per second. Their effect on the ease and accuracy of reading when the signal is weak can be seen in Table 6, and in the photographs, Fig. 12.

4.2.3 Generation of the Radial Trace

The "convertor" circuit, Fig. 10, is interposed between the differential detectors and the corresponding deflector plates of the cathode-ray oscillograph

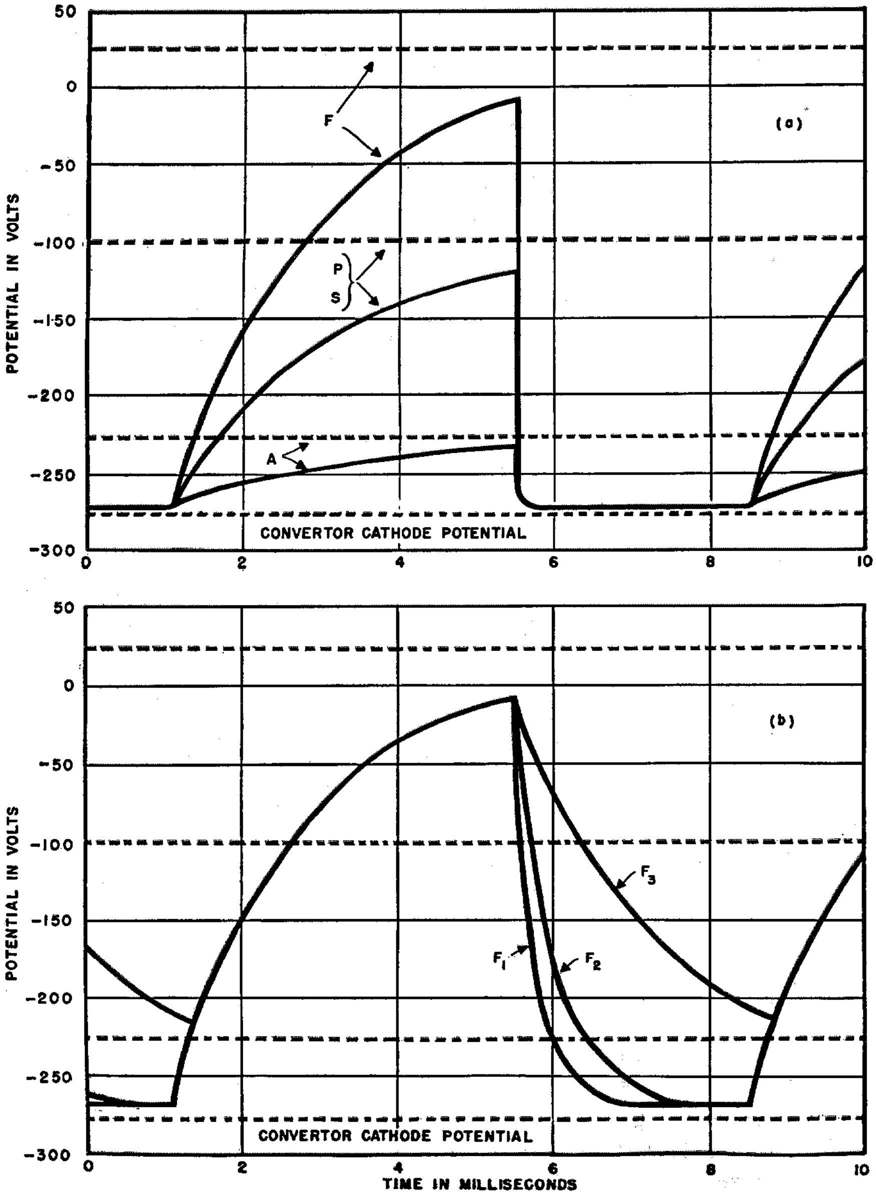


Fig. 11—(*a*) Convertor input potentials and corresponding output waveforms for bearing 000 degrees. (*b*) Repeater output waveforms for various cable time-constants. The waveforms in (*b*) correspond with F in (*a*), and the following time-constants: F_1—0.25 millisecond, F_2—0.5 millisecond, F_3—2.0 milliseconds.

to change the form of the deflecting potentials without affecting their relative amplitudes.

The deflector plates of the cathode-ray oscillograph are connected periodically (at about 135 cycles per second) to a common point by synchronous high-vacuum triode electronic switches controlled by a voltage of rectangular waveform. When the valves conduct, the condensers C_F, C_A, C_P and C_S associated with the F, A, P and S deflector plates of the cathode-ray oscillograph are discharged rapidly through relatively low resistances, and are charged relatively slowly through the high resistances R_F, R_A, R_P and R_S when the valves are cut off. The luminous spot on the cathode-ray-oscillograph screen generates the indicating trace by outward radial movement during the charging part of the cycle, and the discharge "fly-back" is so rapid as to be barely visible at normal brightness. The radial line is straight provided only that the charging time-constants C_FR_F, C_AR_A, C_PR_P and C_SR_S are all equal, a condition which is unaffected by valve characteristics. So long as the valves conduct sufficiently well to bring the trace close to the centre of the screen, their characteristics have no effect on the accuracy of bearing indication. Valve failure is indicated at once by the trace becoming "detached" from the centre of the screen. The charging time-constants are such as to permit the deflector plates to reach about 90 per cent of the steady input potential before discharge occurs.

The steady input potentials and the waveforms of the four output voltages from the convertor circuit, corresponding to the indication of bearing 000 degrees are shown in Fig. 11(*a*). (The voltages applied to the port and starboard deflector plates are identical for this condition.) The relatively long period of quiescence in each cycle ensures the effective marking of the origin of the trace, and facilitates remote indication (Section 4.3).

4.2.4 *Bearing Indicator*

A 12-inch electrostatic cathode-ray tube (CV275) is used as the bearing indicator. Tubes of this type are individually calibrated and read bearings relative to ship's head on a 6-inch-diameter scale attached to the tube face. Bearings relative to true north are read on a separate annular scale of 8.5-inch diameter controlled by the ship's master gyro-compass, using an angular cursor. At the working final anode potential of 1900 volts, the deflection sensitivity of the cathode-ray oscillograph is about 0.4 millimetre per volt for both pairs of plates, the difference being less than 5 per cent.

An auxiliary focusing control is used in an otherwise conventional circuit to bring the radial trace to sharpest focus at the outer extremity, giving it a slight taper. The control is obtained by adjusting the third anode to a potential close to that of the deflector plates at the end of the working (charge) stroke.

The brightness of the trace is sufficient for daylight observation without a viewing-hood; some idea of the general appearance of the display can be obtained from Fig. 12 which shows unretouched photographs of typical bearing indications on strong and weak signals, at 122 megacycles per second. The green fluorescence of the CV275 is not very suitable for photography and a rather long exposure (30 seconds) was used; external interference during exposure has increased the angular width of the trace slightly in some cases.

4.2.5 *Monitoring and Testing Circuits*

These were similar to those of the second experimental model, with the exception of the calibrator. For convenience in manufacture and routine testing, this was constructed as a closed network of fixed non-inductive resistors with 16 tapping-points selected by a rotary stud switch, and corresponding with the cardinal, mid-quadrant and mid-octant bearings.

4.3 REMOTE INDICATION

Remote indication at distances up to 500 feet (cable length) was introduced at a comparatively late stage in the development. Direct parallel connection of corresponding deflector plates of the local and remote oscillographs being impracticable, a cathode-follower "repeater" circuit of four triode-connected CV1091 (EF50) valves was interposed. Transmission of potential variations up to 350 volts necessitated high cathode resistances (50,000 ohms) for power economy, and the limit of signalling distance was set by the discharge time-constant of the line

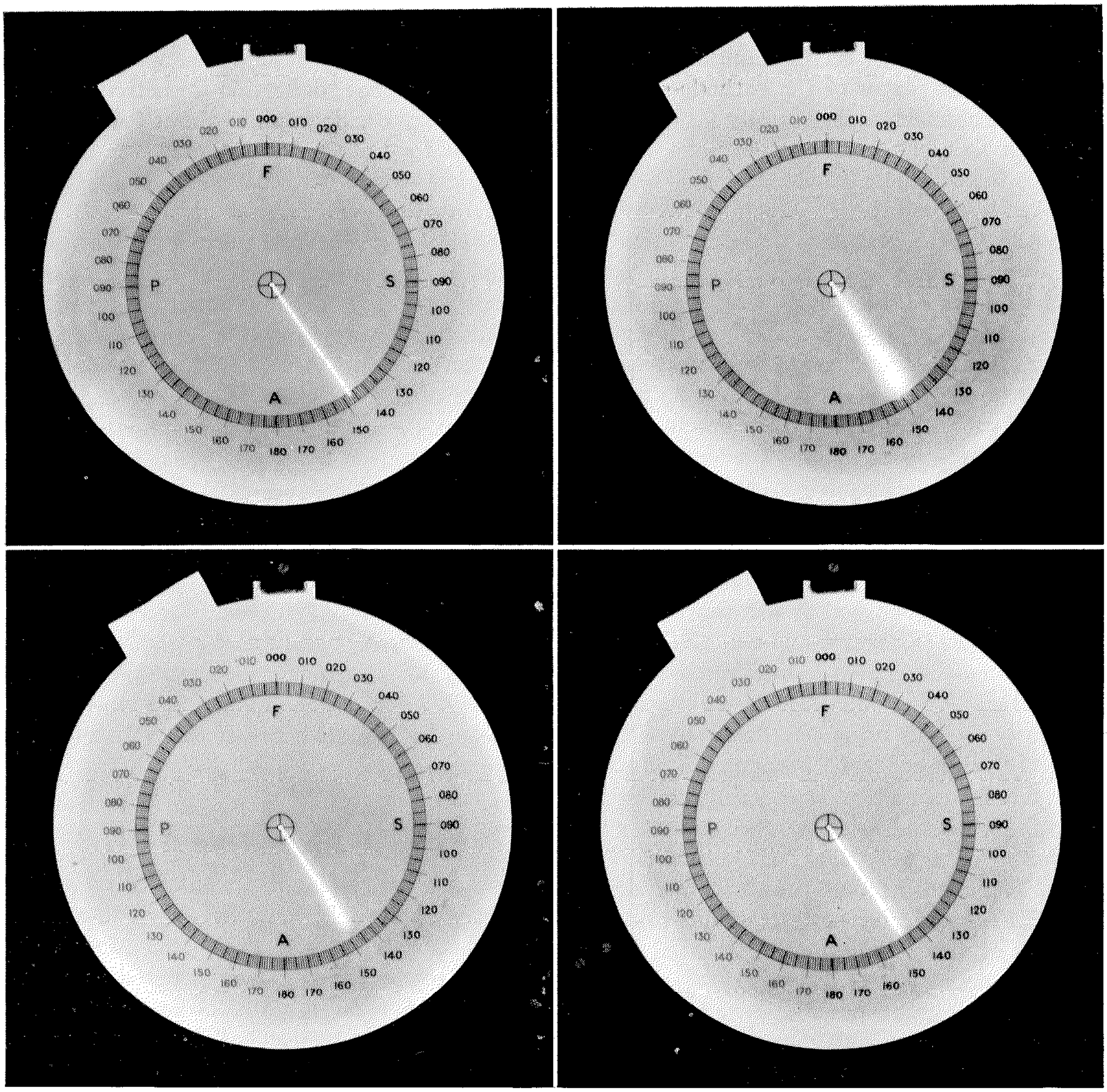


Fig. 12—Typical FV5 bearing indications on cathode-ray tube CV275 (gyro-compass scale and cursor removed). Upper, left—Medium-speed indication, strong signal (about 1 millivolt per metre). Upper, right; Lower, left; and Lower, right—Fast, medium and slow-speed indications on a weak signal (about 6 microvolts per metre).

capacitance/cathode resistance combinations. The curves F_1, F_2 and F_3 of Fig. 11(*b*) show the repeater output waveforms corresponding with cable capacitances respectively below, equal to, and above, the maximum for correct operation. Excessive capacitance causes the radial trace on the remote cathode-ray oscillograph to become "detached" from the screen centre. The beam of the remote cathode-ray oscillograph is cut off during the discharge stroke, which is necessarily slower than on the local tube, and might otherwise cause a prominent and possibly curved "flyback."

Any small relative quadrantal error between local and remote cathode-ray-oscillograph tubes is compensated by a pre-set differential resistive potential divider.

The remote indicator, Fig. 13, includes an oscillograph unit identical with that of the main direction-finder equipment. There are no operational controls other than those required for normal oscillographic adjustments.

4.4 Compensation for Quadrantal Error

Considering the system as a whole, it is clear that although practically all the amplification takes place in a common channel, there remain a number of points in the circuit at which the signals from the two directional-antenna combinations will experience different treatment, resulting in quadrantal error. Compensation for error arising in the low-frequency circuits is made by pre-set adjustment in the differential detector (Section 4.2.2), using a test signal from the calibrator. Any overall quadrantal error is then compensated by differential control of modulator bias, using a vertically polarized test-signal from a source located about 100 feet from the direction-finder antenna system on a mid-quadrant bearing.

In the absence of reciprocal or site errors, it would be sufficient to adjust the indicated bearing of the test signal to the appropriate mid-quadrant reading on the cathode-ray-oscillograph scale. For best accuracy, however, it is generally necessary to take such errors into account by adjusting the indication to a slightly different angle, predetermined as follows:

The antenna system is set successively at the four angles that bring the relative bearing of the source into each of the four mid-quadrant (relative) directions in turn, and after observing the error in each case the mean quadrantal error is computed with due regard for sign, a process which necessarily eliminates the effects of reciprocal and site errors. With normal antenna orientation, the indication on the cathode-ray oscillograph is then re-adjusted to eliminate the computed error, and the indicated bearing is recorded for use in any subsequent re-adjustment. For the most accurate results, the angle is measured for each of the working frequencies.

A method of quadrantal adjustment by direct injection of test signals is under development. While it has the advantage of being independent of site error, there are formidable difficulties in ensuring that the injected signals in each of the three components of the antenna system correspond closely in phase over the whole frequency band with those due to normal reception. In addition, the circuit must be simple and highly stable if it is to serve as a standard of adjustment.

4.5 Performance

Data presented in this section relate to the accuracy and sensitivity of the direction-finder as determined by factory tests and field trials. Figures for accuracy are treated on a statistical basis, since the instrumental error of an automatic direction-finder is a function of both frequency and azimuth.

Adequate information on the performance of an FV5 quipment on a good site is lacking, since the few that have been installed temporarily on land have not been subjected to rigorous trials with co-operating aircraft. The performance on board ship is not of general interest since it does not represent, even approximately, that to be expected on land. It has therefore been considered preferable to quote results obtained with the second experimental direction-finder in the course of extensive field trials carried out with the co-operation of Fighter Command, Royal Air Force, at Tangmere, Sussex. The value of these results is enhanced by the fact that simultaneous trials were carried out on a widely used manual direction-finder of good design, installed on an adjacent site.

The results obtained during these field trials are also of value for the indirect evidence which they afford on the performance of the system in

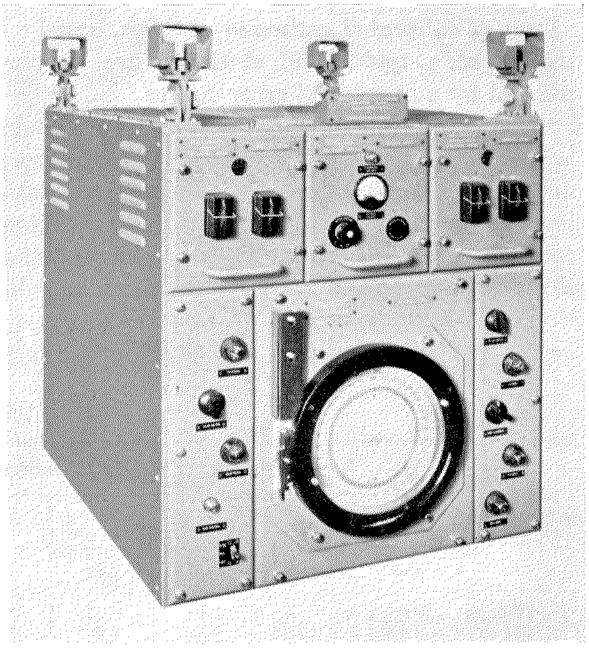


Fig. 13—FV5 remote indicator (for deck-head mounting).

respect of polarization error. No direct quantitative study of this has yet been made.

4.5.1 *Instrumental Accuracy*

The most useful method of expressing the instrumental accuracy of production direction-finders would be by means of a statistical analysis of the overall errors of a series of complete equipments. Unfortunately it is not practicable to test each equipment as a complete installation. In production, each antenna system is tested with one typical direction-finder office equipment on a permanent test-site. All figures relating to the accuracy of antenna systems therefore include the relatively small errors of this equipment. Each production office-equipment, on the other hand, is tested using its own built-in calibrator, and the errors observed include those of the calibrator circuit which is designed for functional tests and is not really accurate enough to serve as an absolute standard. Finally, the direction-finders cannot be tested with the cathode-ray tubes which will be used in subsequent operation. These facts must be borne in mind when considering the results given below.

In this section statistics are given for the accuracy of antenna systems in conjunction with a typical direction-finder office equipment, for the accuracy of remote indication, and for the accuracy of calibrated cathode-ray tubes, based on the maker's tests. The statistics are supplemented by a number of error curves.

Production tests on antenna systems are made at three frequencies after individual modulators and central antenna amplifiers have been subjected to valve ageing for 50 hours under working conditions (to eliminate any early failures) and a short vibration test simulating mast-head conditions in a ship installation (5 minutes vibration at 10 cycles per second, 0.2 inch amplitude). Errors are measured in the four cardinal and four mid-quadrant directions by the method indicated above (Section 4.4). Octantal errors, being calculable, are not normally measured.

The results of routine tests on 20 consecutive antenna systems from the production line are summarized in Table 1, which shows the root-mean-square errors for each frequency, in each of the eight directions of measurement, and in Table 2 which shows for each frequency, (*a*) the probable error, (*b*) maximum errors and (*c*) the percentage of errors exceeding 2 degrees.

TABLE 2

Frequency in Megacycle per Second	Approximate Probable Error in Degrees	Maximum Errors in Degrees		Percentage of Errors Greater than 2 Degrees	Total Number of Observations
		+	−		
100	0.6	2	2.5	0.7	160
122	1.1	2.5	3.5	3	160
150	1.25	3	3	14	160

The distribution of errors for the three frequencies is shown in diagrams (*a*), (*b*) and (*c*) of Fig. 14. Making allowance for the comparatively small number of observations and the fact that errors are only recorded to the nearest 0.5 degree, it is noteworthy that diagram (*a*), for 100 megacycles per second, is a reasonable approximation to the typical probability-curve of purely random errors, such as those of observation. Diagrams (*b*) and (*c*) show quite

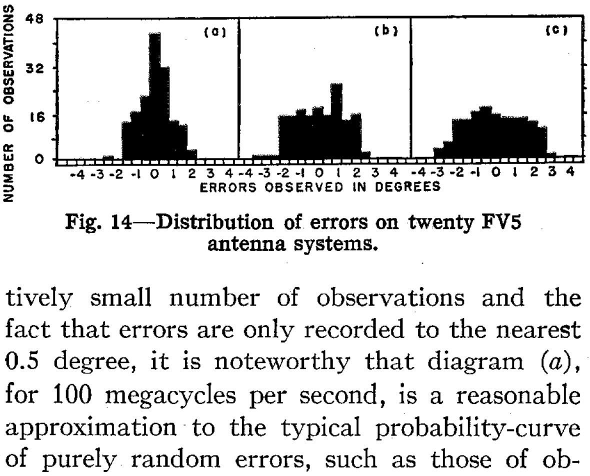


Fig. 14—Distribution of errors on twenty FV5 antenna systems.

TABLE 1

Frequency in Megacycles per Second	Nominal Bearings and Corresponding Errors (Root-Mean-Square Degrees)								Root-Mean-Square Error, All Bearings	Total Number of Observations
	Fore	Starboard			Aft	Port				
	000	045	090	135	180	045	090	135		
100	0.61	0.78	1.03	0.82	0.82	0.95	0.92	0.96	0.87	160
122	0.98	1.23	1.54	1.82	0.46	1.07	1.49	1.61	1.34	160
150	1.06	1.92	1.72	1.46	1.45	1.34	1.34	1.69	1.52	160

different distributions, since at 122 and 150 megacycles per second there are components of reciprocal form, sinusoidal for variation of azimuth. Periodic errors of this type do not give a normal distribution-curve, because the probability that errors will be nearly zero is actually less than the probability that they will be near the maximum for the function. Diagrams (*b*) and (*c*) represent the combined effect of errors that vary sinusoidally with azimuth and of random errors of observation.

The accuracy of the remote indicator system is checked by taking error curves with the same cathode-ray tube placed first in the main indicator position, and then in the remote indicator unit. The maximum relative error permitted is 1 degree. The root-mean-square values of relative error for a group of twelve production equipments are given in Table 3, for each of the sixteen test bearings.

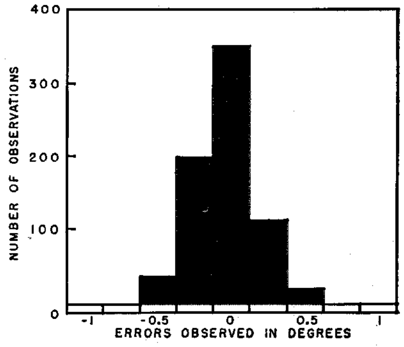


Fig. 15—Distribution of errors on 44 calibrated cathode-ray tubes, CV275.

The accuracy of the calibrated cathode-ray tube Type CV275 is illustrated in the distribution diagram, Fig. 15. This is based on production tests on a run of 44 tubes, measured at angular intervals of 22.5 degrees. The probable error appears to be considerably less than 0.25 degree.

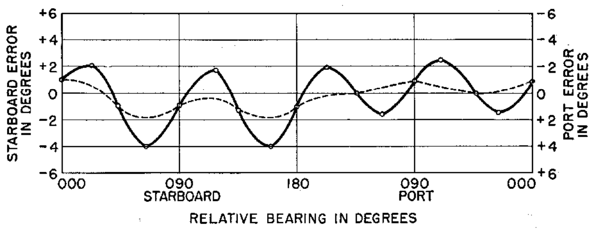


Fig. 16—Typical FV5 instrumental error-curve at 122 megacycles per second. Full line: as taken on uniform scale. Broken line: effect of scale compensation.

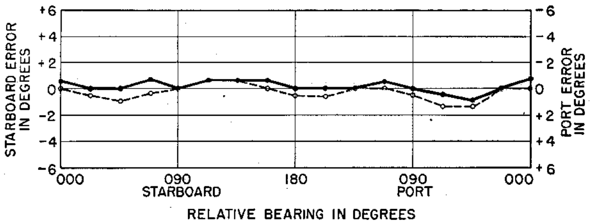


Fig. 17—Apparent errors of typical FV5 indicator circuits. Full line—main indicator. Broken line—remote indicator.

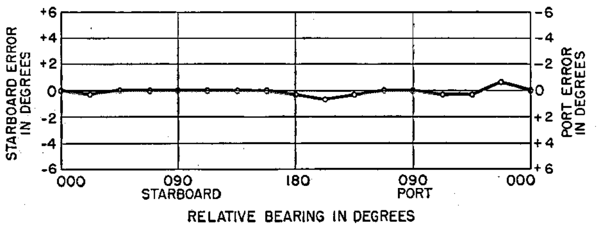


Fig. 18—Apparent errors of indicator circuit used on FV5 antenna test-site.

The overall error-curve for a typical direction-finder is shown in Fig. 16, which illustrates the effect of using a bearing scale compensated for octantal error to the extent of 2 degrees (maximum). Fig. 17 shows the apparent errors of the low-frequency circuit of a typical production office-equipment, including any error contributed

TABLE 3

Nominal Relative Bearing in Degrees	Fore	Starboard							All Bearings, 192 Observations
	000	022.5	045	067.5	090	112.5	135	157.5	
Relative Error, Root-Mean-Square Degrees	0.52	0.38	0.44	0.42	0.13	0.38	0.21	0.46	0.60
Nominal Relative Bearing in Degrees		Port							Aft
		022.5	045	067.5	090	112.5	135	157.5	180
Relative Error, Root-Mean-Square Degrees		0.68	0.58	0.65	0.68	0.60	0.61	0.66	0.31

by the internal calibrator, as explained above. Similarly, Fig. 18 shows the apparent errors of the antenna test-site equipment.

4.5.2 Field Tests on Accuracy

Simultaneous trials were carried out at Tangmere in August–September, 1943, on the second experimental direction-finder described above, and on an Air Ministry manual very-high-frequency direction-finder, Type 2A. They were installed on good average sites about 100 yards apart, and so far as possible bearings were measured simultaneously on the same transmissions.

Test signals were provided by an Avro Anson aircraft flying over a series of "pin-points" on a course embracing the two northern (landward) quadrants, at distances ranging from 25 miles in the east (Brighton), to 55 miles in the west (Wimborne Minster). The heights flown were 2000 to 5000 feet depending on weather conditions and range. Bearings on both direction-finders were observed by Service operators without previous experience of the automatic system. Transmissions were of 5 to 10 seconds duration. All automatic observations have been corrected for the mid-band octantal error of 2 degrees (maximum) as the observations were made on a uniform bearing scale.

Table 4 and the error-distribution diagrams (*a*) and (*b*) of Fig. 19 summarize the results of the five main series of observations, which were carried out at 106 megacycles per second (two series) and at 110, 114 and 120 megacycles per second. The two greatest errors recorded on the manual instrument could not be recorded on the distribution diagram (*a*) due to limitations of space. Test No. 5 was subject to frequent jamming, and the superior performance of the automatic system in this case is noteworthy.

The results as a whole show that the performance of the automatic direction-finder was more consistent, judging from the smaller maximum errors and higher percentage of first-class bearings. However, the best performance on a single series of observations was given by the manual instrument. This result was not unexpected and illustrates the fact that the results obtained with such an instrument depend greatly on the skill of the operator and on the prevailing conditions.

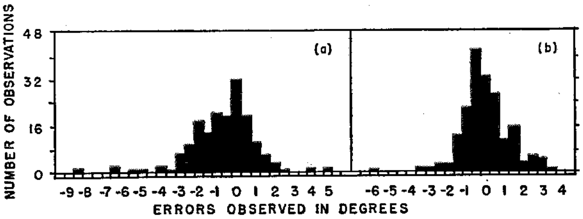


Fig. 19—Distribution of errors for field trials at Tangmere. (*a*) Manual direction-finder, Air Ministry Type 2A (169 observations). (*b*) Second experimental automatic direction-finder (186 observations).

The accuracy of the direction-finders on high-angle signals appeared to be about the same, judging from the results of a number of simultaneous qualitative tests on an aircraft flying directly over the direction-finder site. Automatic indications remained fairly steady up to 45 to 60 degrees elevation, after which increasingly violent fluctuations occurred until at the highest angles one or more complete revolutions were sometimes noted. Corresponding effects were observed on the manual instrument—"flattened" minima, variations too rapid to follow, and failure or reversal of sense indication. The instant of passage of the aircraft over the direction-finder site was easier to detect on the automatic equipment owing to its higher speed of operation.

TABLE 4

Test Number	Frequency in Megacycles per Second	Number of Observations		Root-Mean-Square Error in Degrees		Maximum Error in Degrees		Approximate Percentage of Errors Exceeding 2 Degrees	
		Manual	Automatic	Manual	Automatic	Manual	Automatic	Manual	Automatic
1	106	35	41	1.02	1.93	2.5	6	3	24
2	106	27	31	6.66	1.20	26	3	41	6
3	110	40	40	1.37	1.23	2.5	3.5	5	8
4	114	28	34	2.77	1.35	6.5	3.25	32	9
5	120	40	43	2.37	1.07	10.5	2.5	20	5
All tests	—	170	189	3.20	1.40	26	6	18	11

4.5.3 Effective Direction-Finder Signal-to-Noise Ratio and Deflection Sensitivity

Signal-to-noise ratio at the bearing indicator is a fundamental measure of the performance of an automatic direction-finder, since, for a specified weak signal, it determines the reading accuracy attainable. In this section signal-to-noise ratios are not given explicitly, but in terms of an inversely related quantity, the "fluctuation angle" (±degrees) for signals of known strength.

Deflection sensitivity has little theoretical significance since it can be modified at will by simply changing the overall gain of the common amplification-channel. The gain available in the FV5 equipment is such that all signals that can be identified by speech modulation give a radial trace at least 1.5 inches long (half-scale deflection).

Measurements of fluctuation angle and deflection sensitivity were made on fields of calculable strength. The source of signal was a generator feeding a vertical, elevated, resonant dipole some 20 feet from the direction-finder antennas and at the same height above ground-level. An independent experiment showed that free-space propagation could be assumed without sensible error and the field was calculated from the relation

$$\mathcal{E}_V = 60I/d,$$

where $\mathcal{E}_V$ = field strength in volts per metre,
I = antinode current in amperes, calculated from generator output-voltage, and
d = distance from dipole in metres.

Mean figures quoted in the tables were based on three sets of measurements, made with two generators of the same type and a third of an entirely different type. The spread of individual observations did not exceed ±15 per cent about the mean figures given.

TABLE 5

Frequency in Megacycles per Second	106	122	150
Field Strength in Microvolts per Metre for 1.5-Inch Trace	6.5	4	2.5

Table 5 shows the mean field strengths required for a 1.5-inch radial trace at three frequencies distributed through the band.

Table 6 shows the approximate fluctuation-angle for the three speeds of indication, for a trace length of 1.5 inches, corresponding to the field strengths of Table 5. For a given trace-length the fluctuation angle was roughly independent of frequency, within the limits of experimental error.

TABLE 6

Indication Speed	Approximate Time-Constant of Differential Detector in Seconds	Approximate Fluctuation-Angle in Degrees
Fast	0.03	±9
Medium	0.3	±3
Slow	1.0	±1

4.5.4 Field Tests on Effective Range

An indication of the effective range of the FV5 direction-finder is given by the results of some tests carried out on the second experimental model in the course of the field trials at Tangmere briefly described in Section 4.5.2. Owing to unfavourable weather it was not possible to obtain more than a limited number of observations, mainly at 120 megacycles per second. The aircraft carried a transmitter rated to deliver 5 watts and flew at heights of 6000 to 7000 feet; the direction-finder antenna-system was mounted at a height of 24 feet. Bearings were observed in the north-west quadrant and the trace length was recorded as the aircraft reported position at preselected points, identified visually. Trace lengths of 1 to 2 inches (about half-scale deflection) were observed on 8 occasions at an average range of 89 miles, and of 2.5 inches or more on 17 occasions at ranges averaging 75 miles. The results were rather erratic, including a deflection as small as 1.5 inches at only 77 miles, and a full-scale deflection at 98 miles; this was doubtless due in part to the nature of the intervening country and perhaps to the aspect of the aircraft.

The figures suggest a half-scale-deflection range of about 80 miles at 120 megacycles per second for an aircraft at 6000 feet, and a few observations made at 106 megacycles per second suggest a corresponding range of about 70 miles at the lower frequencies in the band 100–125 megacycles per second.

With automatic gain-control in use the deflection-to-range characteristic is very flat up to ranges of 65 to 75 miles (aircraft at 6000 feet), but falls off extremely sharply for ranges exceeding that at which half-scale deflection is obtained. This is due to the combined effect of the falling-off in field strength according to an inverse square (or higher) law and the square-law rectification of weak signals.

In range, as in accuracy, the results obtained with the manual direction-finder were found to depend greatly on the skill of the operator, but on the average the maximum effective range was about the same for both systems.

5. *Very-High-Frequency Automatic Direction-Finders for Land-Station Service*

The Admiralty D/F Outfit FV5 is being adapted for use on land, at direction-finder stations associated with civil and military aerodromes. The immediate requirements for this service are remote indication with speech monitoring and facilities for adjusting quadrantal balance in the control tower, while the main direction-finder equipment works unattended on an open site which will generally be at least half a mile from the main aerodrome buildings. Taking a longer view, full remote operation with frequency selection from the control tower is desirable for economy in staff, while simultaneous multi-channel working and interconnection of local and remote equipment over normal telephone lines are equally desirable for economy in plant and transmission circuits.

5.1 DIRECTION-FINDER TYPE P.V.1—A

Adapted to meet the immediate requirements stated above, this is essentially the same as the FV5 except that the bearing repeater circuit has been modified to permit remote indication up to about a mile from the main direction-finder equipment, or up to nearly three miles with the sweep frequency of bearing indication reduced from 135 to 50 cycles per second.

5.1.1 Remote Indicator Circuit

Reference to curve F_3 of Fig. 11(*b*) (Section 4.3) shows that to increase the range of remote indication of the FV5 system it is necessary either to accelerate the discharge of the line capacitances during the "flyback" periods, or to reduce the sweep frequency. For distances up to about one mile the former method was adopted,[13] the discharge being accelerated by electronic switches (triode-connected CV1060) shunted across the cathode-follower load resistances and controlled in parallel by a voltage derived from the radial sweep circuit (Section 4.2.3). By this means the line capacitance which could be discharged effectively at 135 cycles per second was increased from 0.01 microfarad to about 0.15 microfarad. Triode-connected beam tetrodes CV345 were about four times more effective in reducing the discharge time-constant but did not allow a corresponding increase in line capacitance, which at a sweep frequency of 135 cycles per second was limited by the current-carrying capacity of the cathode-followers to about 0.25 microfarad. This corresponds to about one mile of light-gauge lead-covered and rubber-insulated power cable of 660-volt grade.

5.1.2 Stability of Direction-Finder Circuits in Unattended Operation

For stability in unattended operation the most significant adjustments are those of differential detector centering and phasing, and overall quadrantal balance. The latter depends mainly on constancy of relative gain between the antenna modulators.

Stability was investigated in a series of 6 tests of 8 hours and one of 32 hours on a standard Admiralty FV5 equipment. Measurements on the significant adjustments made at hourly intervals yielded similar results in all tests. The results of the final 32-hour test were as follows:

Differential detector-centering measured in terms of cathode-ray-oscillograph deflection was constant within ±0.75 millimetre throughout the period. Variations of this order might lead to errors up to 1 degree on a weak signal giving half-scale deflection, but as the centering may be corrected at the remote indicator their effects may be discounted.

Differential detector-phasing was constant within ±1 degree throughout the period. The apparent variations observed were no greater than the limits of experimental error in adjustment, and were clearly so small as to have a completely negligible effect on the quadrantal balance.

Hourly variations of overall quadrantal error are plotted on the graph of Fig. 20 for the three routine test frequencies 100, 122 and 150 megacycles per second. Large variations of alternating-current line voltage are known to affect the quadrantal error, but there is no apparent correlation in the case of the small variations recorded during the test, also plotted on Fig. 20. The differences between the mean quadrantal errors are due to the fact that the balance adjustment was left untouched throughout the period. The limits of variation for the three frequencies were approximately ±0.5 degree about the mean values, ignoring the observations of the first two hours.

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[13] British Patent No. 590261.

The variation of quadrantal balance with time is almost entirely due to small changes of uncertain nature associated with the modulator units and central antenna amplifier, and further tests were made to determine the time taken by these elements to reach steady operating conditions. Three pairs of modulators were tested, including the pair used in the 32-hour test. All modulators were tested in association with each pair of directional antennas (6 tests in all), and quadrantal error was measured at intervals of 5 minutes until quasi-stability was reached. It was found that by a coincidence the combination and arrangement used for the 32-hour test gave by far the worst results, requiring 60 minutes to reach a quadrantal error within 1 degree of the final mean error. In the other five tests the corresponding times were 10 minutes or less.

5.2 Direction-Finder Type P.V.1—B

This represents a current development of the FV5 direction-finder and goes further toward meeting the general requirements already stated. The direction-finder antenna system can be arranged to cover one of the bands 100–118, 118–132 or 132–156 megacycles per second without re-tuning the output circuit. It feeds two receivers tuned to different frequencies, one of which may be reserved for distress calls. The receivers have separate speech-channels in continuous operation, and when a signal is received on either channel, separate visual-warnings are given by neon lamps both in the direction-finder station and at the remote indicators. Under standby conditions, direction-finder signals from both channels are passed to the common indicator-circuit. If signals are received on both frequencies simultaneously the operator can switch the bearing indicator to one channel or the other. Up to three remote indicators can be used, and the system can be controlled either at the direction-finder station or from any of the remote indicators. Obstruction warning lamps can be mounted on the sides of the antenna box when the direction-finder station is located close to an airport.

Remote indication is obtained by the transmission of direct-current deflecting potentials from the differential detectors, using a cathode-follower circuit and a radial-sweep generator at the remote control-point. Leakage resistances as low as 30,000 ohms have no effect on the accuracy of bearings, and automatic audible warning is given of any leakage capable of affecting the accuracy. High-speed indication (Section 4.2.2) is possible over 20 miles of a special dry paper-insulated cable which also carries the speech and remote-control circuits.

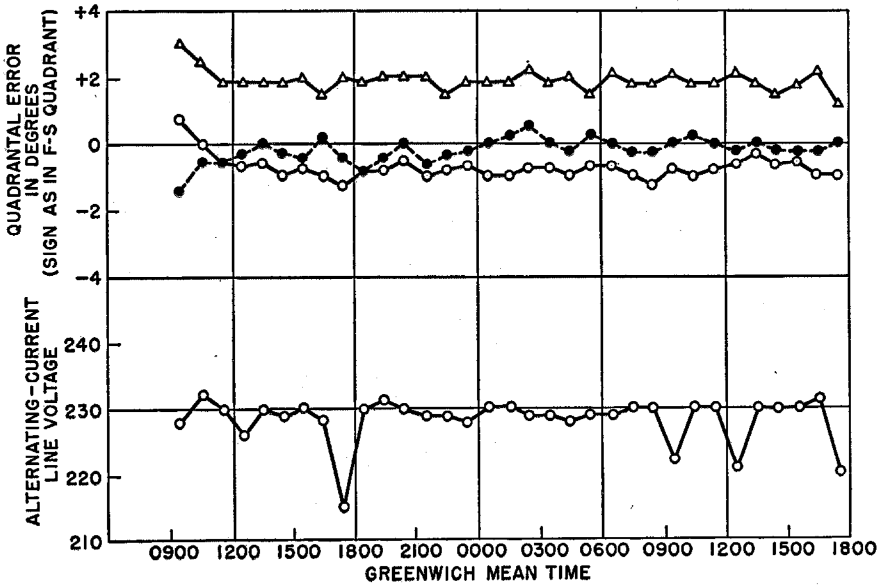

Fig. 20—Quadrantal balance stability of FV5 direction-finder in continuous operation. Operating frequencies in megacycles per second: triangles—150, circles—122, and dots—100.

5.3 Future Developments

These are expected to include full remote-control and indication together with simultaneous multi-channel operation. Transmission of bearing information over a single circuit is considered essential, both for accuracy and for economy, so retaining the basic feature of the system. The problem has been solved in principle, but discussion of the subject is beyond the scope of the present paper.

6. *Special Applications of the System*

It is proposed to use a direction-finder similar to the P.V.1—B in a projected navigational system described in a companion paper,[14] in which the position of an aircraft is determined in polar co-ordinates relative to a single base-station, the radial co-ordinate being determined by a co-operative phase-comparison method using continuous waves. In another application the P.V.1—A direction-finder is being used as a means of identifying the indications given by a primary radar system with plan-position display. It is only necessary here to give a brief account of the latter application.

6.1 Trace Identification in the Air-field Control Radar, British Air Ministry Mk. IIIx

The A.C.R. Mk. IIIx is a primary radar system giving plan-position indications and is used in the control of aircraft within a radius of about 20 miles from an airport. Display units are installed in the control tower, and for effective operation it is essential to be able to associate the radar traffic pattern with specific aircraft. The addition of direction-finder indications to the radar display has proved the simplest available method, and although it must occasionally fail when two aircraft appear temporarily on the same line of bearing, a high percentage of identifications is immediately successful.

The P.V.1—A direction-finder station equipment is installed in a small hut about 100 yards from the radar antenna-system, at a point close to the centre of the airfield. The radar and direction-finder indications are obtained on separate oscillograph tubes and combined optically in a special display-unit (Air Ministry

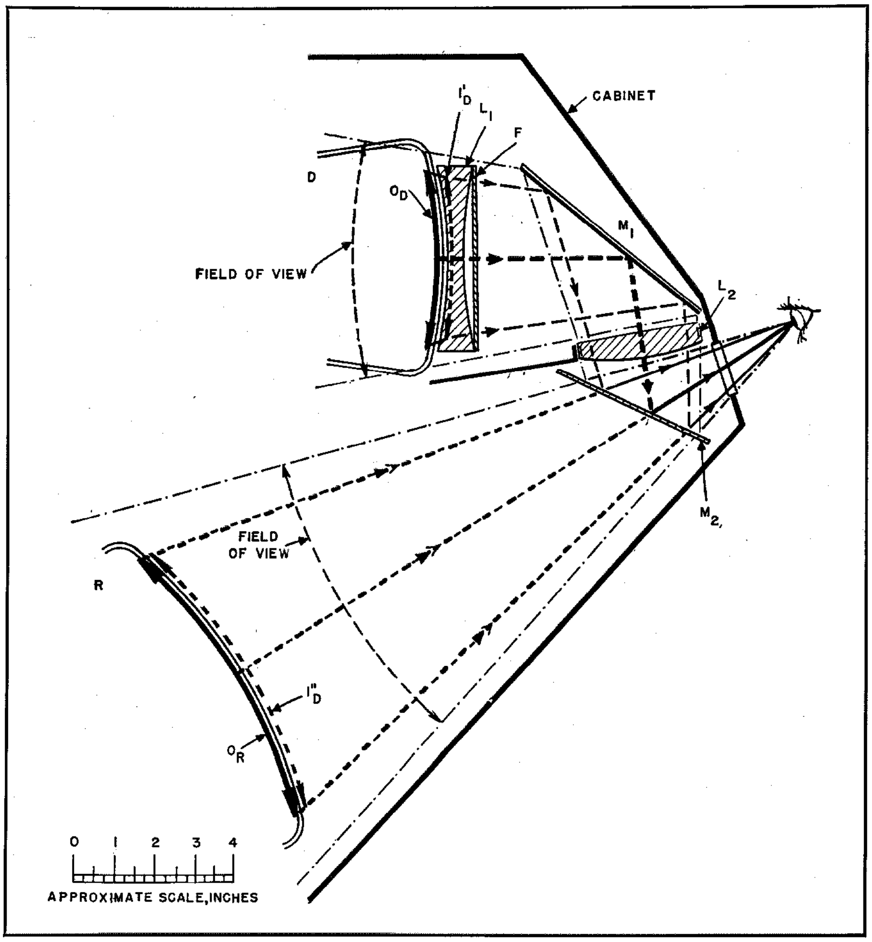

Fig. 21—Optical system of display unit DU74. D, direction-finder oscillograph. R, Radar oscillograph. L_1, Correcting lens. F, Blue colour filter. M_1, Polished steel reflector. L_2, Magnifying lens. M_2, Plate-glass reflector incorporating yellow colour filter. O_D, direction-finder trace. I'_D, Intermediate image of O_D, curvature corrected. I''_D, Final image of O_D. O_R, Radar image. (I''_D is normally adjusted to coincide with O_R.)

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[14] R. F. Cleaver, "Note on a Short-Range Radio Position-Finding System Using Modulated Continuous Waves," *Journal of the Institution of Electrical Engineers*, v. 94, Part IIIA, p. 984; 1947: and *Electrical Communication*, v. 25, pp. 363–372; December, 1948.

Type DU74). While any system of separate tubes has obvious disadvantages compared with one employing a common oscillograph on a time-sharing basis, there are notable compensating advantages which are apt to be overlooked. First, the use of separate oscillographs makes it possible to display the direction-finder bearings on a screen of short persistence, thus avoiding the confusing afterglow effects which would otherwise occur during brisk direction-finder activity. Further, the direction-finder tube can have a screen fluorescence of low intensity and contrasting colour, giving clearly distinguishable bearing indications without obliterating weak radar signals. Finally, complete electrical separation of the direction-finder and radar circuits simplifies maintenance and reduces the risk of simultaneous failure of both parts of the system.

The principle of the optical system used in the display unit is illustrated in Fig. 21, while Fig. 22 shows the corresponding side view of the display unit with covers removed.

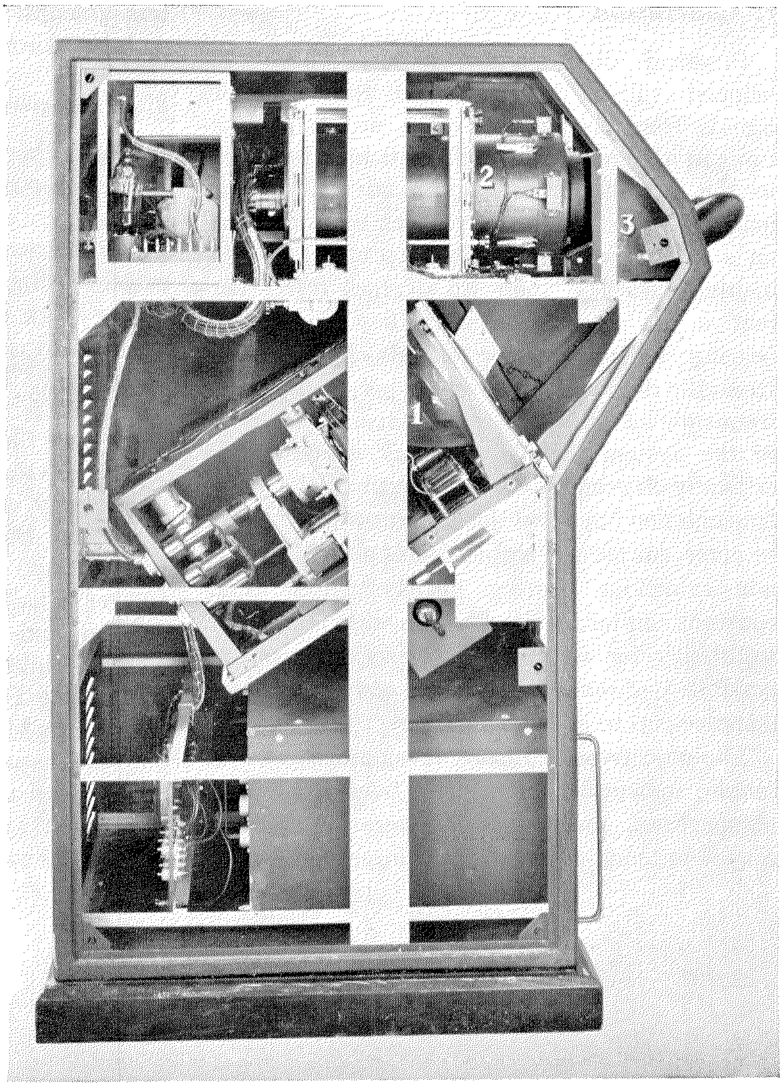

Fig. 22—Display unit DU74 with side cover removed. (See optical diagram, Fig. 21.) (1) Cathode-ray tube, plan-position indicator. (2) Cathode-ray tube, direction-finder. (3) Optical unit.

Referring to Fig. 21, R and D represent in cross-section the radar and direction-finder tubes, R having a long-persistence screen with yellowish phosphorescence, and D a screen giving blue fluorescence of short persistence. M_1 is a metallic reflector and M_2 is an unsilvered plate-glass partial reflector incorporating a yellow colour-filter. The radar screen is observed directly through M_2, which largely suppresses the fluorescence or "flash," and the direction-finder indication is viewed by double reflection at M_1 and M_2. Back-surface reflection of the direction-finder trace at M_2 is eliminated by the combined action of the yellow filter in M_2 and a complementary blue filter, F, in front of the direction-finder tube. The magnifying lens L_2 makes it possible to use a relatively small (7-inch) direction-finder tube and to mount it further forward than would otherwise be possible. The bi-concave contact lens L_1 corrects the apparent curvature of the image of the screen of the direction-finder tube, and virtually eliminates parallax over the whole of the field of view. All mechanical adjustments necessary to bring the two screens into the correct relative position and orientation are provided on the mountings of the direction-finder tube.

7. Conclusion

Practical direction-finders have been developed utilizing the principle of amplifying the signals from fixed, orthogonal antenna structures in a common channel. Instruments for use in the very-high-frequency band 100 to 150 megacycles per second have been manufactured on a production basis, with a probable instrumental error less than 1.5 degrees (after correction for octantal error). They give useful bearing indications for signals having field strengths down to 7 microvolts per metre at the frequency of minimum sensitivity. Advantages of the system include accuracy which is not critically dependent on the length of the single transmission line from the antenna system to the receiver, simple and stable circuit adjustments, a non-ambiguous bearing display whose time-constant can be adjusted within wide limits, and suitability for unattended operation with the possibility of remote indication and control over distances up to at least 20 miles.

The objectives of future development include remote indication and control over trunk telephone lines, multi-channel operation from a single antenna-system, and improvement in instrumental accuracy, possibly by the use of wide-aperture antenna-systems.

8. Acknowledgments

The author wishes to record his appreciation of the co-operation of many colleagues in the work described in this paper, and regrets that personal acknowledgment is impossible in most cases. Specific reference must, however, be made to the work of the team headed by Mr. L. J. Heaton-Armstrong, responsible for the engineering of the FV5 equipment, and to the part played by Mr. C. W. Earp, whose advice and encouragement were throughout invaluable.

The development was initiated on the responsibility of Mr. C. E. Strong, Chief Radio Engineer, Standard Telephones and Cables, Limited, and most of the later work was carried out on behalf of the Ministry of Aircraft Production and the Admiralty Signal Establishment. Thanks are due to the Board of Admiralty and to the Directorate of Communications Development, Ministry of Supply, for permission to publish the paper.

The statistical information on the accuracy of the CV275 (Fig. 15) is reproduced by courtesy of Edison Swan Electric Company, Limited.

Note on a Short-Range Radio Position-Finding System Using Modulated Continuous Waves*

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THE PAPER describes a position-finding system by which the polar co-ordinates of any suitably equipped aircraft can be determined at a single base station using phase-comparison range-finding apparatus and a very-high-frequency automatic direction-finder in conjunction with conventional communication equipment.[1]

An experimental model is described in which azimuth and range are displayed on a single cathode-ray oscillograph, which may be installed in an airport control tower. Range is indicated without ambiguity up to 100 nautical miles, and there is no azimuthal ambiguity. So far, the model has been used primarily to demonstrate the principles of the system, in conjunction with an aircraft fitted with experimentally modified British Air Ministry very-high-frequency communication equipment. The results of a limited number of tests suggest that with airborne equipment specifically designed for use with the system, the probable error in range measurement should be less than 1 mile. Azimuthal accuracy is not here discussed, but in a companion paper it is shown that the probable instrumental error of the direction-finder is less than 1.5 degrees, implying a probable lateral error less than 2.5 miles at the maximum non-ambiguous range of 100 nautical miles.

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1. Introduction

Radio position-finding systems for aircraft fall into three classes characterized by their dependence on the measurement of two azimuths, of a range and an azimuth, or of two ranges. The present paper is concerned with a new system of the second class in which the position of the aircraft is measured at a ground-based station. Continuous-wave transmissions are used in the measurement of both range and azimuth, and the system is cooperative in the sense that a signal transmitted from the ground is repeated back from the aircraft. In these basic respects the system resembles the earlier German "*Y*" system (British nomenclature: "Benito"), though it differs radically in circuit technique and in the manner in which the positional information is presented.

In the new system, an automatic direction-finder of a type described in a companion paper[2] is used to indicate the bearing of the co-operating aircraft by means of a radial unidirectional trace on the screen of a cathode-ray oscillograph. The manner in which the radial trace is generated makes it particularly suitable as a time base for the measurement of range, and it is to this fact that the present system owes its origin.

So far, work on the system has been exploratory. An experimental model has been constructed for use either as a short-range navigational aid, or to facilitate the control of aircraft movements in the outer approaches to an airport, indication being given directly in the control tower. The position of any one co-operating aircraft can be established in a few seconds in the absence of significant interference, which, if present, is apparent to the operator.[3] The experimental model operates on continuous-wave or telephony-modulated signals transmitted in the very-high-frequency band 100–150 megacycles per second, these signals being used also to

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* Reprinted with minor additions from *Journal of the Institution of Electrical Engineers*, v. 94, Part IIIA, n. 16, pp. 984–989; 1947.

[1] The system has become known under the synthetic name "*Condar*" derived by abbreviation from "continuous-wave direction- and range-finder.

[2] R. F. Cleaver, "The Development of Single-Receiver Automatic Adcock Direction-Finders for Use in the Frequency Band 100–150 Megacycles per Second." *Journal of the Institution of Electrical Engineers*, v. 94, Part IIIA, n. 15, pp. 783–795; 1947: also *Electrical Communication*, v. 25, 337–362; December, 1948.

[3] The effects of interference in the direction-finder used are discussed in the companion paper previously cited.

identify the co-operating stations by the normal communication procedure. A demonstration of the experimental model was given for delegates to the Provisional International Civil Aviation Organization at the Royal Aircraft Establishment, Farnborough, England, in September, 1946.

In the experimental form of the system described in the paper, both direction and range are measured by means of signals which can be transmitted and received by normal communication equipment with comparatively small modifications, so that a minimum of additional equipment is required in the aircraft. The service offered is equivalent to that obtainable from a group of direction-finding stations arranged for triangulation and providing automatic indications on a common cathode-ray oscillograph at a central plotting station. In cases where such an arrangement would be a practicable alternative, the new system offers considerable economies in staff, equipment, lines and buildings. Moreover, since the ground equipment can be concentrated in a very small area if necessary, the system has the further advantage that it can operate in areas where the triangulation method would be uneconomic or inapplicable, e.g., at key points on long air routes over undeveloped or desert country, or in the neighbourhood of island bases on transoceanic routes respectively.

These advantages are obtained at the expense of (*a*) modifications to the aircraft very-high-frequency communication equipment, involving an increase in weight which might be as little as 2 pounds in a favourable case, but which might be as much as 12 pounds, and (*b*) the use of separate frequencies for air-to-ground and ground-to-air communication. The necessity for two frequencies is not an unmitigated disadvantage, since it improves the efficiency of communication considerably, and is sometimes favoured on this score alone. Further, in those areas where position-finding by triangulation is impracticable, it is unlikely that the allocation of a second frequency would present any difficulties.

2. *Principle*

An aircraft receives from the ground station a carrier wave of frequency f_1 modulated at a frequency f_R, and simultaneously transmits a second wave, frequency f_2, carrying modulation of the same frequency and phase, derived from the received signal. The polar co-ordinates of the aircraft position are determined at the ground station by an automatic direction-finder tuned to the aircraft transmitting frequency f_2, and by a range-finder which operates by measuring the phase difference ϕ between the modulations on the signals transmitted from, and received at, the ground station. The distance d is indicated directly, and is related to the phase difference in degrees by the equation

$$d=\frac{\phi c}{720 f_R}, \tag{1}$$

where the same units of length are used for d and for c, the velocity of light.

The direction-finder gives the bearing indication on the cathode-ray-oscillograph screen in the form of a radial trace generated by exponential outwardly directed sweeps of the electron beam, synchronized with the modulation so as to provide a time base on which the position of the aircraft may be displayed and its range measured. The measurement is made with reference to an electronic scale marked on the time base, or by means of an electronic "cursor" controlled by a calibrated phase-shifter. These methods were adopted because the physical length of the time base depends to some extent on the received signal strength.

2.1 CHOICE OF CARRIER FREQUENCIES

The association of exploratory work on the system with direction-finder developments in the frequency band 100–150 megacycles per second resulted in the use of frequencies in this band for both f_1 and f_2. In fact, however, the only theoretical limitation in the choice of frequencies is that, within the working range of the direction-finder, both the signals should travel by paths not differing appreciably in length from the distance between the aircraft and the ground station. This admits considerable freedom of choice in respect of the ground-to-air frequency f_1, even though availability of direction-finder equipment at present restricts the choice of f_2 to frequencies in the particular very-high-frequency band mentioned.

2.2 Choice of Modulation Frequency

This is essentially a compromise between the conflicting requirements of high accuracy on the one hand and freedom from ambiguity on the other. If instability of circuit phase-shift and limitations in the measuring technique result in a probable error of $\delta\phi$ degrees in the measurement of the phase shift in transmission, and if the accuracy required is specified by a probable error δd, the modulation frequency is given by

$$f_R = \frac{c\delta\phi}{720\delta d}. \qquad (2)$$

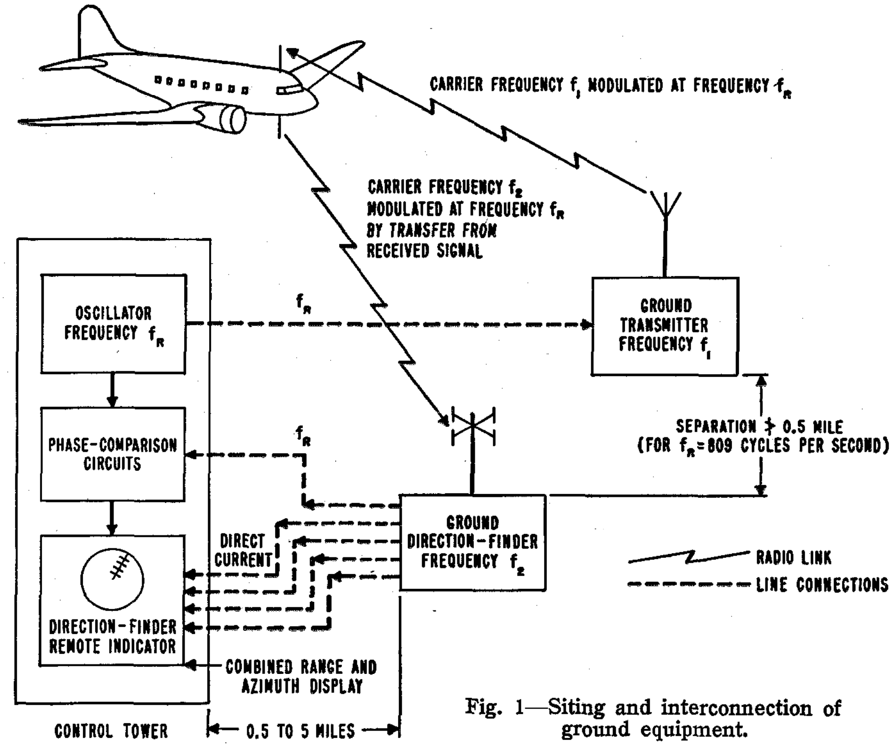
CARRIER FREQUENCY f_1 MODULATED AT FREQUENCY f_R
CARRIER FREQUENCY f_2 MODULATED AT FREQUENCY f_R BY TRANSFER FROM RECEIVED SIGNAL
OSCILLATOR FREQUENCY f_R
f_R
GROUND TRANSMITTER FREQUENCY f_1
PHASE-COMPARISON CIRCUITS
f_R
SEPARATION ≯ 0.5 MILE (FOR f_R=809 CYCLES PER SECOND)
GROUND DIRECTION-FINDER FREQUENCY f_2
DIRECT CURRENT
DIRECTION-FINDER REMOTE INDICATOR
RADIO LINK
LINE CONNECTIONS
COMBINED RANGE AND AZIMUTH DISPLAY
CONTROL TOWER
0.5 TO 5 MILES

Fig. 1—Siting and interconnection of ground equipment.

Since $\delta d/\delta\phi$, and hence f_R, are fixed by the accuracy specified, the first ambiguity is encountered at a distance D, where

$$D = \frac{360\delta d}{\delta\phi}. \qquad (3)$$

In the present system, it should be possible to maintain $\delta\phi$ at 4 degrees or less,[4] and from (3) the corresponding probable error in range measurement should not exceed about 1 per cent of the maximum non-ambiguous range. If the latter were arranged to exceed the greatest very-high-frequency communication range currently attainable, the accuracy would be insufficient at the more usual operating distances. By way of compromise, the modulation frequency has been chosen to give no ambiguity up to 100 nautical miles, a conservative figure for the maximum range of the direction-finder for an aircraft at an altitude of 10,000 feet. When there is a possibility that the co-operating aircraft may be in communication at a distance exceeding 100 nautical miles, it will generally be possible to resolve ambiguity by consideration of received signal strength, the height and estimated position of the aircraft, and other relevant factors.

The modulation frequency corresponding to a phase shift of 360 degrees at a range of 100 nautical miles is 808.9 cycles per second (taking $c = 2.9978 \times 10^{10}$ centimetres per second) and for practical purposes this is rounded off to 809 cycles per second. The range accuracy obtainable with a phase shift of 3.6 degrees per nautical mile should be adequate over most of the normal service area of the direction-finder, and generally consistent with the lateral accuracy attainable.

For the measurement of very short ranges, considerably greater accuracy would be appropriate, and would require the use of a second, higher modulation frequency, a lower frequency being retained for the resolution of ambiguity. In the German "*Y*" system (Siemens model) frequencies of 300 cycles per second and 3000 cycles per second were used in this way,[5] corresponding to phase shifts of 1.34 and 13.4 degrees per nautical mile, respectively. The lower frequency resolved ambiguity up to 270 nautical miles (500 kilometres), and it is claimed that an operational accuracy "within 0.5 mile" was obtained.

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[4] This assumes that the strength of the carrier wave received from the aircraft is great enough to ensure something like full-scale deflection on the oscillograph. This assumption is justified for all ranges and altitudes at which the aircraft could be identified by the speech modulation on carrier waves received from it.

[5] In the later "Graetz," "Rechlin" or "Becker-Gestell" system, a frequency shift (3000–3300 cycles per second) was used, with equivalent result.

3. Siting and Interconnection of Ground Equipment

A convenient arrangement of ground equipment is illustrated in Fig. 1. The operating position may be in the control tower of an airport, the equipment here consisting essentially of a direction-finder remote indicator, a source of modulating tone for range determination, and circuits for range calibration and remote control of the direction-finder and transmitter.

The transmitting and direction-finder stations are located on adjacent sites which may be up to several miles from the control tower. In general, the spacing between the former will be several hundred yards, the minimum distance being set by considerations of direction-finder site error. Any spacing whatever renders the position determination subject to what may be termed a "parallax" error, of which the law of variation depends upon the point selected as the origin of coordinates, and of which the magnitude increases with the spacing. Errors in range measurement are minimized by taking the origin at the mid-point of the line joining the direction-finder and transmitting antennas, while azimuthal errors are clearly a minimum if the direction-finder antenna system is taken as the origin. In both cases the maximum parallax error is equal to half the spacing between the direction-finder and transmitting antennas, and should preferably be small compared with the probable instrumental error of the system. In the particular form of the system described in the paper, the spacing could be as much as 0.5 mile without introducing significant error.

Land-line connections convey the modulating tone from the control tower to the transmitting station, and the output at modulation frequency from the direction-finder receiver back to the tower, for phase comparison. Phase shifts in these lines, and in the ground equipment circuits, are compensated at the operating position by a preset adjustment which can be checked at any time in a few seconds. Bearing information is transmitted to the remote indicator in the form of four steady potentials, over a separate 4-wire line, and the quadrantal balance of the direction-finder may be checked and corrected by the use of a signal generated at the direction-finder site and controlled from the tower. In addition to the lines mentioned, conventional circuits are

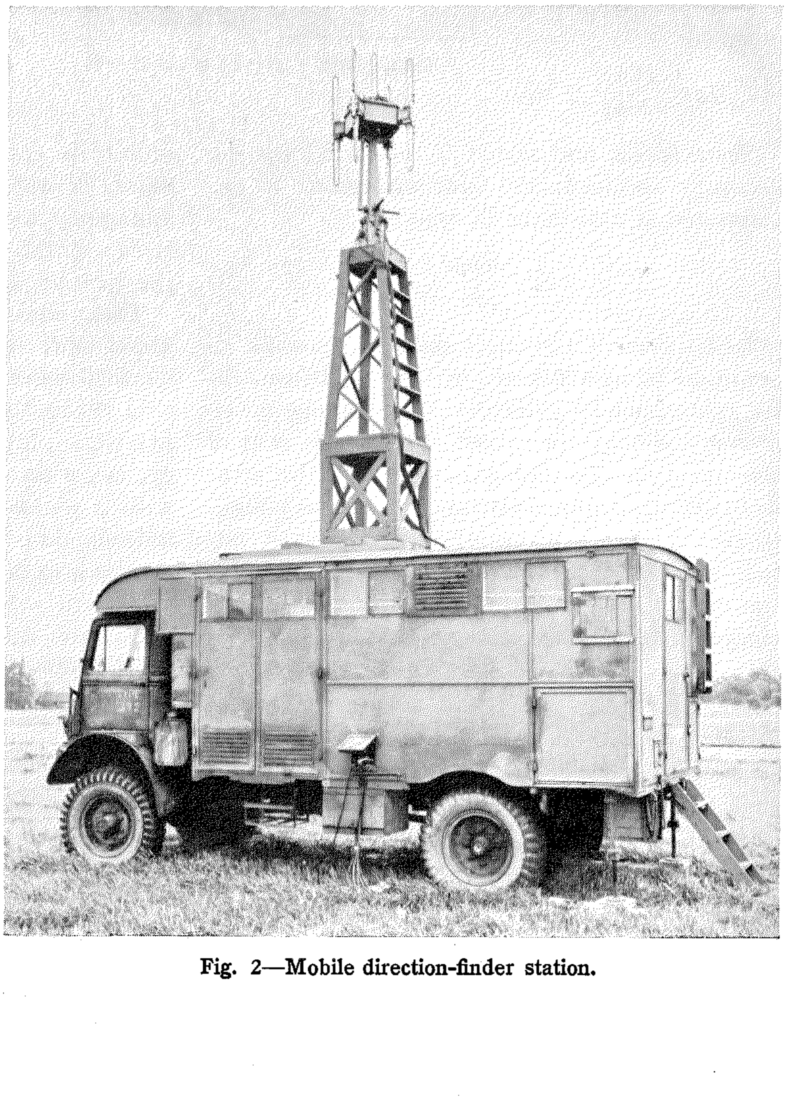

Fig. 2—Mobile direction-finder station.

provided for remote control of the transmitter and direction-finder.

As an alternative to the arrangement described above, it might be advantageous in some cases to use a separate ground receiver for range determination. For maximum accuracy in position finding, the direction-finder station would then be located midway between the transmitting and receiving antennas.

4. *An Experimental Model*

The model of the system used for the Provisional International Civil Aviation Organization demonstrations employed a single modulation frequency of 809 cycles per second. With the exception of the experimental control-tower apparatus, it was assembled from the most readily available current Service equipment, modified when necessary, and, in the case of the air-borne very-high-frequency communication set, not particularly suitable for the purpose.

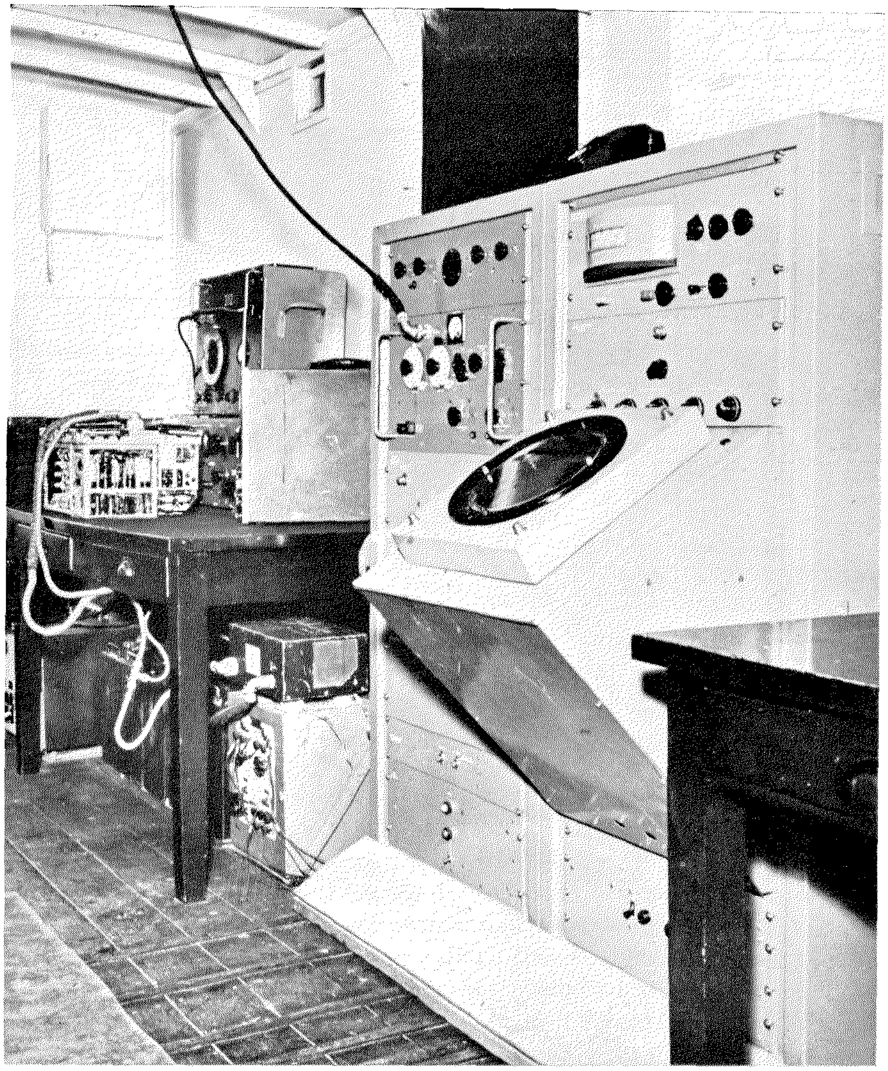

Fig. 3—Experimental control-tower equipment. Aircraft communication equipment TR-1143A on test bench in background.

The whole of the ground equipment was installed on an open site close to the aerodrome boundary. The transmitter, Air Ministry type T.1131, and the direction-finder, Admiralty D/F Outfit FV5, were housed in motor vehicles placed about 250 yards apart. The direction-finder vehicle is shown in the photograph, Fig. 2, with the antenna system mounted on a wooden tower at a height of about 22 feet. The transmitter was provided with a vertical quarter-wave aerial with horizontal counterpoise, at a height of 25 feet. The control tower equipment, illustrated in the foreground of Fig. 3, was installed in a trailer about 80 yards from the direction-finder vehicle. The interconnection was as described in the previous section.

For demonstration purposes, an Avro Anson aircraft was fitted with an Air Ministry very-high-frequency communication equipment, type TR.1143A, experimentally modified as described in Section 4.3. This equipment can be seen in the background of Fig. 3, set up on a test bench for lining-up.

Fig. 4 shows the complete circuit in block

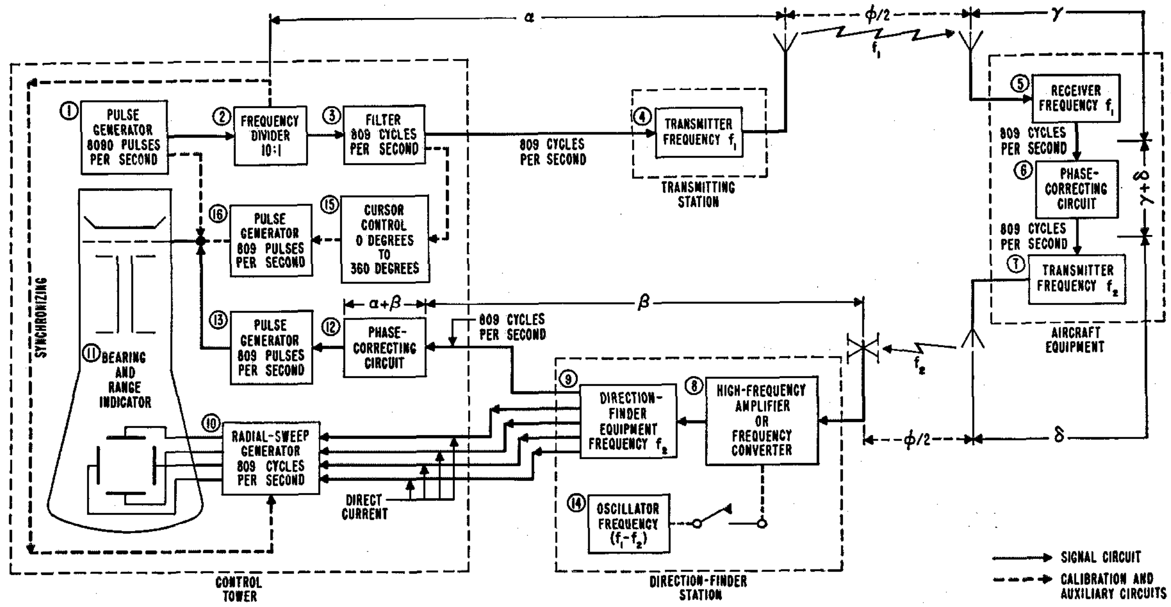
α
ϕ/2
γ
f_1
① PULSE GENERATOR 8090 PULSES PER SECOND
② FREQUENCY DIVIDER 10:1
③ FILTER 809 CYCLES PER SECOND
809 CYCLES PER SECOND
④ TRANSMITTER FREQUENCY f_1
TRANSMITTING STATION
⑤ RECEIVER FREQUENCY f_1
809 CYCLES PER SECOND
⑥ PHASE-CORRECTING CIRCUIT
γ + δ
809 CYCLES PER SECOND
⑦ TRANSMITTER FREQUENCY f_2
AIRCRAFT EQUIPMENT
SYNCHRONIZING
⑯ PULSE GENERATOR 809 PULSES PER SECOND
⑮ CURSOR CONTROL 0 DEGREES TO 360 DEGREES
α+β
β
⑬ PULSE GENERATOR 809 PULSES PER SECOND
⑫ PHASE-CORRECTING CIRCUIT
809 CYCLES PER SECOND
f_2
⑪ BEARING AND RANGE INDICATOR
⑨ DIRECTION-FINDER EQUIPMENT FREQUENCY f_2
⑧ HIGH-FREQUENCY AMPLIFIER OR FREQUENCY CONVERTER
ϕ/2
δ
⑩ RADIAL-SWEEP GENERATOR 809 CYCLES PER SECOND
DIRECT CURRENT
⑭ OSCILLATOR FREQUENCY (f_1-f_2)
CONTROL TOWER
DIRECTION-FINDER STATION
SIGNAL CIRCUIT
CALIBRATION AND AUXILIARY CIRCUITS

Fig. 4—Block schematic circuit diagram.

schematic form, only the experimental sections being indicated in any detail.

4.1 Signal Circuit

The modulating tone at 809 cycles per second originates in the control tower, where the equipment includes a master oscillator running at 8090 cycles per second, driving a pulse generator (1) followed by a 10:1 frequency divider (2) and filter (3) whose sinusoidal output at 809 cycles per second is sent over a land-line to the transmitter (4) to modulate a carrier wave of frequency f_1. This is received in the aircraft (5), and the 809-cycle-per-second component from the receiver second detector passes through a phase-correcting circuit (6) to the transmitter (7), to modulate the radiated signal, frequency f_2. This is picked up by the direction-finder antennas, amplified (8) and applied to the direction-finder equipment (9). Output from the latter includes a component at 809 cycles per second whose phase relative to the modulation on the signal radiated from the ground is a measure of the range, and four steady deflecting potentials whose relative magnitudes are determined by the direction of arrival of the signal. The 809-cycle-per-second and direct-current components of the signal are conveyed over suitable land-lines to the control tower, together with speech output from the receiver. In the control tower the steady potentials are converted into exponential saw-tooth form at a frequency of 809 cycles per second in the radial-sweep generator (10), and are then applied to the deflector plates of the cathode-ray-oscillograph bearing and range indicator (11) to produce a radial trace on the screen. The component at 809 cycles per second passes through the phase-correcting network (12) to a pulse generator (13) whose output is applied to the grid of the cathode-ray oscillograph (11), producing a bright spot on the radial trace at a distance from the centre depending on the range of the aircraft. The outward radial sweep is synchronized with the modulation on the ground transmitter by an auxiliary connection from the 10:1 frequency divider (2) to the sweep generator (10).

The duration of the sweep can be adjusted to be just over one-half of a cycle of modulation at 809 cycles per second, so that ranges up to 50 nautical miles (180 degrees of phase shift) can be displayed. By reversing the phase of the modulation on the ground transmission, the apparent phase shift is reduced by 180 degrees so that ranges from 50 to 100 nautical miles can be measured on the same time base. This arrangement effectively doubles the length of the exponential range scale, which for the normal full-scale deflection averages 1.8 millimetres per nautical mile, and has a maximum value of about 2.5 millimetres per nautical mile at the central origin of the time base.

The positional accuracy of the system depends upon two pre-set adjustments which are made in the control tower, and may be checked in a few seconds at any time. These adjustments, described in Sections 4.1.1 and 4.1.2, affect range and azimuthal accuracy respectively.

4.1.1 *Zero Range Adjustment*

Unavoidable phase shifts in the ground transmitting and receiving circuits (α and β, Fig. 4) are corrected in the network (12). The adjustment is made with the direction-finder receiver temporarily retuned to the ground transmitting frequency f_1 by means of a frequency changer formed by amplifier (8) in association with the beat oscillator (14) which may be switched on from the control tower. The range indication is adjusted to one-half of the distance between the ground transmitter and the direction-finder station.

4.1.2 *Quadrantal Balance Adjustment*

The quadrantal balance of the direction-finder may be checked and adjusted if necessary, using a test signal on a mid-quadrant bearing, provided by an oscillator under the control of the operator.

4.2 RANGE CALIBRATION CIRCUIT

On the experimental model, two methods of measuring range on the radial time base are provided. In the first, positive pulses from the generator (1) are applied to the grid of the cathode-ray oscillograph and produce bright spots on the radial time base, constituting an electronic scale with intervals of ten nautical miles. The range of an aircraft can then be obtained rapidly by interpolation with sufficient accuracy for many puposes. The second method of measuring the range is by means of an "electronic cursor." An output from the 809-cycle-per-second filter (3) is passed by way of a calibrated phase shifter (15) to a pulse generator (16) whose output is applied to the grid of the cathode-ray oscillograph. The corresponding bright spot on the radial trace constitutes the "cursor," and is brought into coincidence with the aircraft indication by the phase shifter (15), whose calibration (in miles) then indicates the range. A zero adjustment not shown in Fig. 4 is used for initial alignment of the phase shifter against the electronic scale previously described. The phase shifter is continuously adjustable from 0 to 360 degrees with uniform calibration, and is similar in principle (see Section 7) to the angular calibrator used on the direction-finder.

The scale and cursor indications are distinguished from that due to the aircraft by an artifice not shown in Fig. 4. The calibrating

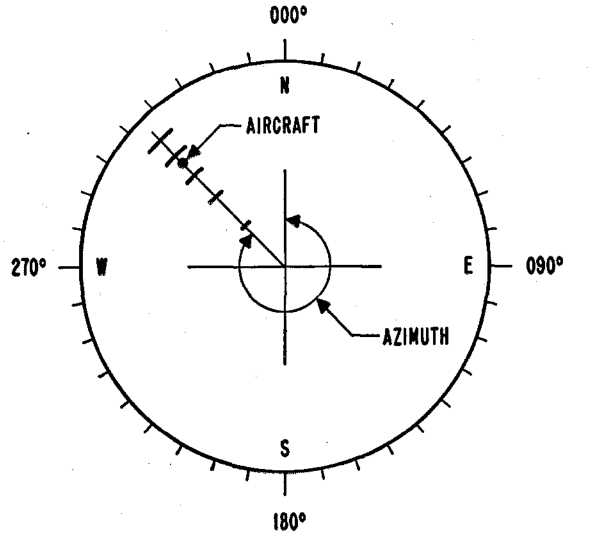

Fig. 5—Range and azimuth display, with electronic scale of 50 nautical miles. Aircraft at 36 nautical miles.

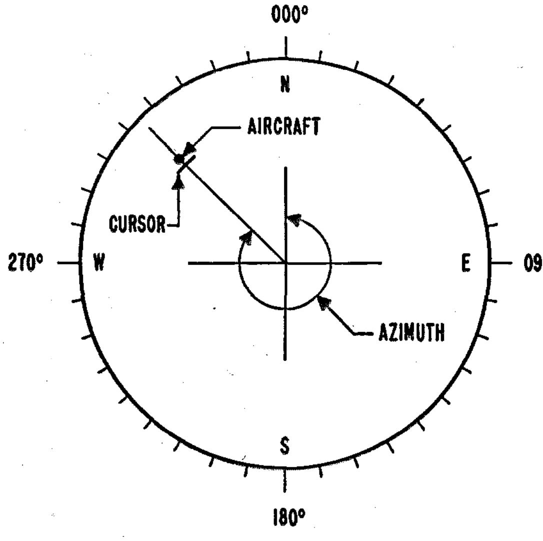

Fig. 6—Range and azimuth display, with electronic cursor. Aircraft at 36 nautical miles. Cursor at 33 nautical miles.

pulses are applied to the cathode-ray oscillograph control grid during alternate sweeps of the radial time base, and the pulse due to the aircraft is applied only during the intervening sweeps. An angular oscillation, of about 5 degrees amplitude and at a frequency of about 200 kilocycles per second, is imposed on the radial time base during calibration sweeps only, thus converting the scale and cursor spots into short arcuate lines. A negative bias applied to the cathode-ray oscillograph grid during calibration sweeps extinguishes the radial trace except at the instants when pulses occur. Figs. 5 and 6 illustrate the forms of indication obtained with the alternative methods of measuring the range. In Fig. 6, the cursor trace is shown displaced by about 3 miles from the "aircraft" spot, which is at a point corresponding to 36 miles.

4.3 Aircraft Equipment

The TR.1143A aircraft equipment was modified for demonstration purposes as follows:

A. Two quarter-wave antennas were provided to permit simultaneous transmission and reception, in place of the common antenna and change-over relay arrangement normally used.

B. A small control box was added, containing a key which, overriding the normal control box, arranged the circuit for simultaneous transmission and reception on two predetermined frequencies. At the same time, output from the receiver was connected to the modulation amplifier of the transmitter.

C. A simple phase-correcting circuit [(6) in Fig. 4] was introduced between the receiver and the modulation amplifier.

D. The modulation amplifier was modified by the application of automatic gain control to the first stage, derived from the modulator output voltage.

The modifications involved an additional weight of about 2.5 pounds, using components of normal size. It may be noted here that in cases where the normal very-high-frequency equipment is fundamentally unsuitable for simultaneous transmission and reception, it would be necessary to provide a separate miniature receiver, weighing perhaps 10 pounds.

Modulation-frequency phase shifts in the aircraft equipment (γ and δ in Fig. 4) are corrected in the adjustable network (6) interposed between receiver and transmitter. This is pre-set during circuit alignment on the ground, by applying to the receiver a signal modulated at 809 cycles per second from a generator, and equalizing the phases of the generator modulating voltage and the corresponding voltage appearing at the output terminal of the transmitter modulator.

4.4 Experimental Results

The Farnborough installation was used almost entirely for demonstration to the Provisional International Civil Aviation Organization and other delegations, under conditions that precluded systematic recording of significant observations. One of the few recorded tests was made to determine the maximum effective range of the system with respect to an aircraft at a height of 2000 feet and its results may be of interest.

The aircraft flew on an outward radial track, ranges obtained by direct visual observation being reported at intervals of 5 statute miles; simultaneously, measurements were made by the electronic cursor method, and in Fig. 7 the errors are shown plotted against reported range. The observation at zero (plan) range has been corrected for the height of the aircraft. Circles in Fig. 7 correspond to the errors of mean measured ranges, while the thick vertical lines show the extent of the "spread" due to fluctuation noise. The maximum effective range for the conditions of the test can be seen to be about 35 miles, at which distance a "spread" of about ±5 miles and an error of −1.5 miles were obtained. At a range of 40 miles, the signal-to-noise ratio was so low that speech was scarcely intelligible in either direction, showing that the useful ranges for

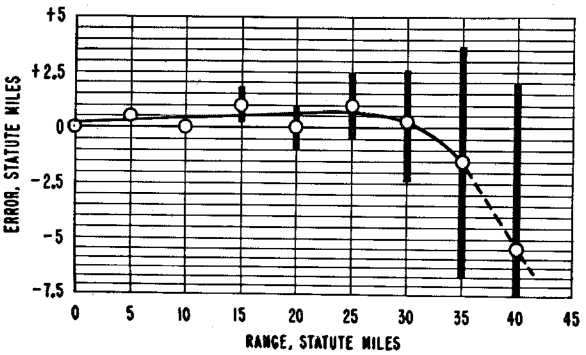

Fig. 7—Range test on an aircraft at a height of 2000 feet. f_1 = 100.8 megacycles, f_2 = 110 megacycles.

speech communication and distance measurement were about the same. Up to 30 miles, the maximum overall error did not exceed 1 mile, and the distribution of errors suggests a reading accuracy not worse than ±0.5 mile. It should be noted that the instrumental accuracy depends fundamentally on the frequency of the master oscillator, nominally 8090 cycles per second, which on the experimental equipment was not crystal-controlled and was not checked after leaving the factory. Observations at the shorter ranges suggest that the oscillator may have been running some 2 to 3 per cent fast.

During the demonstrations, the greatest errors observed were due to variations in modulation-frequency phase shift in the airborne equipment, which in the course of a flight of several hours sometimes amounted to the equivalent of 2 to 3 miles. It should perhaps be emphasized that the circuit of the TR.1143A was standard in all essential respects, and no attempt had been made to improve it in respect either of magnitude or stability of phase shift. Reference to the figures given in Table 1 for a particular equipment shows that, although the resultant phase shift is not very great, there are substantial shifts in individual sections of the circuit, a factor which militates against stability.

TABLE I

TR.1143A COMMUNICATION EQUIPMENT

Measured Phase Shifts at 809 Cycles per Second (Phase Reversals Neglected)

Section	Phase Shift in Degrees
Receiver High- and Intermediate-Frequency Amplifiers	− 6*
Receiver Second Detector (Filter Circuit)	−42
Transmitter Modulation Amplifier	+30
Complete Equipment	−18

* Modulation envelope shift.

Subsequent tests showed the phase shift in the transformer-coupled modulation amplifier to be particularly undesirable, since it was dependent to a marked degree on the input voltage to the motor generator from which the equipment drew its power. The average effect was to introduce error at the rate of about 0.1 mile for each 1 per cent drop in battery voltage from the value used during alignment on the ground. Since in flight the equipment was supplied from an auxiliary battery, and was generally in continuous operation for two or three hours, the observed variations in phase may well have been largely due to supply-voltage changes. In any communication equipment specifically designed or modified for regular operation with the system it is clear that special attention must be paid to low-frequency transformer-coupled stages.

5. *Conclusion*

The work described demonstrates the practicability of a direct-reading ground-based position-finding system utilizing continuous-wave very-high-frequency communication and direction-finder equipment. By the use of the simplest form of range-finding circuit with a single modulation frequency chosen to give no range ambiguity up to 100 nautical miles, it would be reasonable to expect a probable error of one mile or less in the radial co-ordinate, assuming the circuit of the airborne communication equipment to be designed for high stability of phase shift at the modulation frequency of 809 cycles per second. The probable lateral error in position-finding, with the direction-finder used, is not likely to exceed 2.5 miles at a range of 100 nautical miles.

6. *Acknowledgments*

The use of the author's radial-trace bearing indicator as a time base to give a combined display of range and azimuth was the result of a suggestion by Mr. C. W. Earp, to whom thanks are also due for advice and criticism relating to circuit technique. The author wishes to express his appreciation of the co-operation of Mr. T. D. Gray, who carried out a large part of the experimental work. A number of other colleagues rendered valuable assitance, in particular Messrs. P. Sothcott, F. G. Cockerill and T. J. Cox.

The demonstration to the Provisional International Civil Aviation Organization delegations was made possible by the co-operation of the Ministry of Supply, whose help is gratefully acknowledged.

The photographs, Figs. 2 and 3, were taken during the Provisional International Civil Aviation Organization demonstrations at Farnborough and are reproduced by courtesy of the Controller of H.M. Stationery Office. Crown Copyright is reserved.

7. *Appendix—Principle of the Linear Phase Shifter*

The principle of the calibrated phase shifter used in the electronic cursor circuit (Section 4.2) is illustrated in Figs. 8(*A*) and 8(*B*). In Fig. 8(*A*), *AB* . . . . *GH* . . . . represents a closed network of *n* equal uniform resistance elements arranged in one plane as a regular polygon with centre *O*. An *n*-phase generator is connected to the junction points of the network in progressive phase sequence, as indicated. *OQP* is a uniform resistive radial contact arm, grounded at *O* and capable of rotation about this point, making continuous contact with the network, compared with which its resistance is very high. The reference phase for the circuit is that at point *A*, and the angular position of the contact arm is θ, measured clockwise from *OA*.

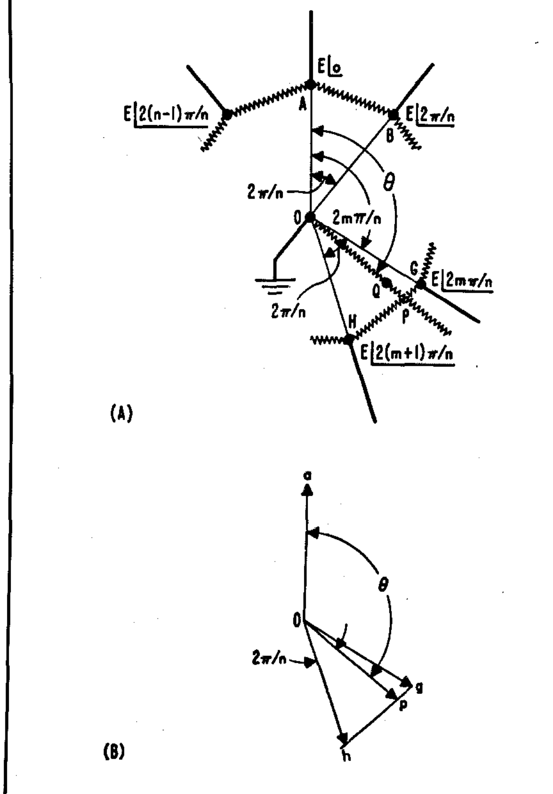

Fig. 8—Principle of linear phase shifter. (*A*) Polygonal arrangement of *n* resistance elements. (*B*) Vector diagram.

Fig. 8(*B*) shows the relevant part of the vector diagram, in which *Oa*, *Og*, *Oh*, represent the voltages to ground at *A*, *G* and *H*. If *Op* is drawn to meet *gh* at *p*, and if $gp/gh=GP/GH$, then *gp* represents the voltage between *G* and *P*, and *Op*, the vector sum of *Og* and *gp*, represents the voltage to ground at the point of contact *P*. The phase of the voltage at this point is given by *aOp* and is clearly equal to θ, the angular position of the contact arm. By inspection of the diagrams, it will be seen that both the voltage between point *P* and ground, and the effective resistance of the contact arm, vary with θ according to the same law, and therefore the current through the contact arm is a constant. If the output voltage e is taken from a point *Q* on the arm, within the inscribed circle of the polygonal network, and if $OQ/OA=k$, it is clear that $e=kE\underline{/\theta}$. The calibration of the phase shifter is therefore uniform, and its output voltage is independent of phase angle.

The use of two resistive elements in rubbing contact is undesirable in a practical realization of the principle, and can be avoided very simply. In the experimental equipment described in the paper, $n=4$ and the network reduces physically to a square, with balanced voltages in quadrature applied between diagonally opposite points.

Design of an Ionization Manometer Tube*

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AN IONIZATION MANOMETER tube of new design is discussed, having a high sensitivity substantially independent of variations in the electrode potentials. The published measurements of efficiency of ionization are compared and used to predict the sensitivity of the gauge on different gases.

Curves of a function used in the calculation of the sensitivity are applicable to any design, and it is shown that secondary emission is an important factor in determining the slope of the anode characteristic.

Errors from leakage and photoelectric currents, gas take-up and gas heating are considered.

• • •

Just as in 1905, when the production of tantalum lamps required a sensitive vacuum meter, the metal-filament lamp became a Pirani Gauge,[1] so later the triode assisted in its own manufacture by becoming an ionization manometer.

The step of using the positive ion current flowing to the negative grid of an operating triode containing gas for the measurement of pressure is usually ascribed to Buckley[2] and Hausser, Ganschwindt and Rukop.[3] Fogel,[4] however, finds that it was described first by Baeyer.[5] In addition to these originators, Dushman and Found[6] in their 1921 study of the gauge state:

> Some time ago Dr. Hull suggested to the writers that a gauge may be based on the measurement of the amount of positive ionization produced by the electron stream. After some preliminary experiments had been carried out with a gauge based on this suggestion, a paper on the same subject appeared by O. E. Buckley.

The gauge rapidly displaced Knudsen's absolute manometer and the molecular gauge of Langmuir in vacuum-tube manufacture, and has been used widely up to the present time. It is highly sensitive, simple, readily degassed and, at vacua better than 1 microbar, the ion current varies linearly with pressure. The variation in sensitivity between different gases will be discussed later. These advantages are attained fully only when the gauge tube is suitably designed, although most amplifying tubes will operate as pressure indicators.

The brief description in Table 1 does not do justice to the designs described in references 4 and 6. Characteristics other than *S*, *A* and *C* are important.

1. Development

Early during 1946, it became necessary to design a gauge tube to replace the triode type 4101 (Fig. 1), which previously was used as a manometer in the manufacture of transmitting valves. The electrode system resembles that described by Jaycox[8] (Table 1) and the sensitivity is comparatively high. To suit existing equipment, the same or a higher sensitivity was required in the new design. Also, it was hoped to achieve this with a much less expensive construction. At first, several simple assemblies were tried. These were tested by comparison with a known manometer by sealing all tubes to a short large-diameter header connected by a length of small tube to a normal pumping system consisting of liquid-air trap, mercury-diffusion and

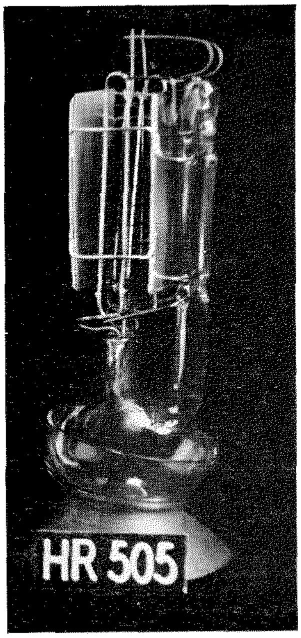

Fig. 1—Electrode system of the 4101 triode.

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[*] Published in abridged form in *Proceedings of the Institution of Radio Engineers* (Australia), v. 8, pp. 14–19; April, 1947, and pp. 4–10; May, 1947. This device is the subject of Australian Patent Application 6735, September 18, 1946.

[1] Numbered references will be found on page 385.

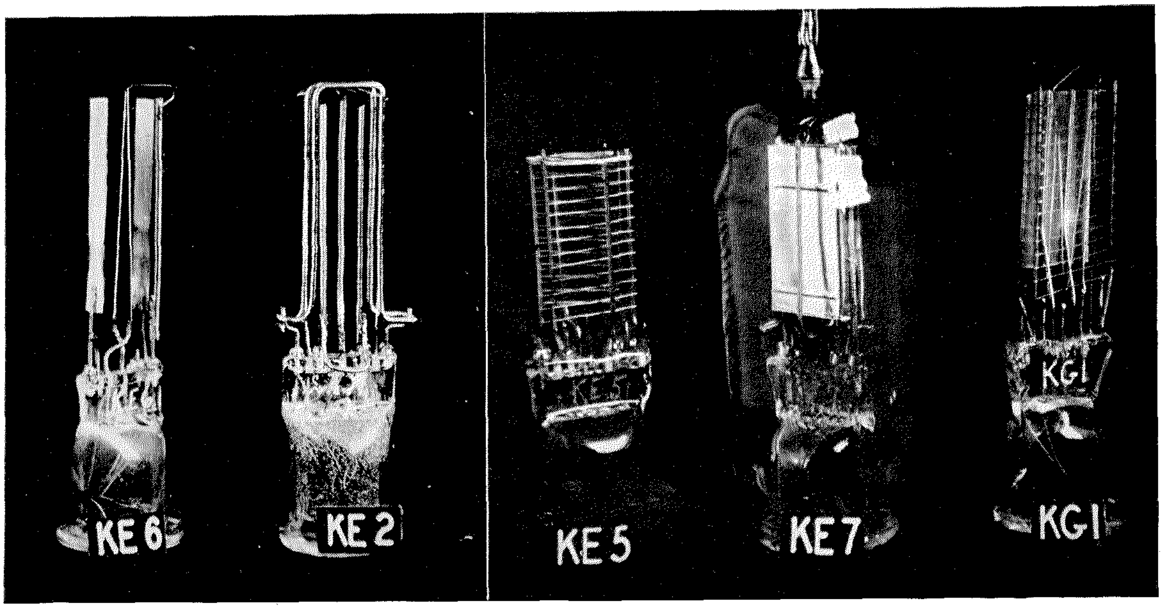

Fig. 2—Experimental manometer designs. At 125 volts anode potential, the measured sensitivities were: KE6 with strip electrodes, $S=1.7$; KE5 with aligned plane grids, inner positive, $S=0.35$, outer positive, $S=3.7$; KE7, plane triode with grid positive, $S=3.0$; KG1 with plane grid, plane plate, and central filament, grid positive, $S=3.2$, plate positive, $S=1.2$.

backing pumps. After an oven bake of one hour and general degassing, gas was released in the header and simultaneous readings were taken on two manometer equipments of the electronic control type. Some of the first experimental designs are shown with characteristics in Fig. 2. These were difficult to degas electronically and were insensitive.

Consideration of the reasons for their failure suggested the single-sided design (KF1, Fig. 3) having a curved reflector which was more successful ($S=6.3$). Moreover, on electron bombardment it degassed readily.*

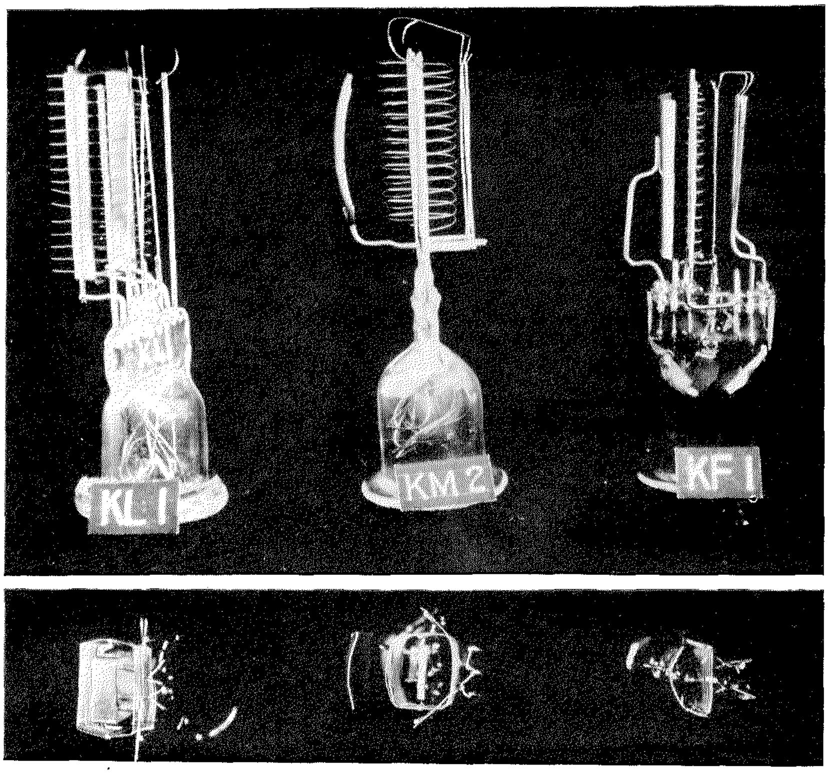

Fig. 3—Designs having higher sensitivities than those shown in Fig. 2.

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* If the area of the electrodes is too large, it is difficult to bring them to red heat by electron bombardment without exceeding a safe filament emission because gas pressures of around 0.5 microbar may limit diode voltage drop to 20–40 volts. If the area is too small, on bombardment the bulb may not be warmed sufficiently above its normal operating temperature in applications where no oven bake is possible.

1.1 "Long-Path" Design

KL1, shown in Fig. 3, represents an entirely different construction in which a rectangular nickel frame (collector) is surrounded by a positive grid. The objective here is to keep the electrons, over their whole path through the grid, near the velocity at which maximum ionization takes place. For these tubes, with a grid 0.45 inch wide, S reached 14 and this principle was adopted in all later tubes. While the optimum proportions for this design were being sought, it was found that under some conditions of bombardment, with comparatively high gas pressure, local heating produced in the glass wall could produce cracks. This trouble is caused when the secondary emission coefficient of the glass wall exceeds unity; allowing the surface to rise to some positive potential. The

TABLE 1

Published Information on Ionization-Manometer Design

Described By	Construction	Design Features	Sensitivity S*	Anode Characteristic A†	Collector Characteristic C†
Buckley[2] 1916	Three Vee filaments in parallel planes 5 millimetres apart; collector central.	Electrodes may be degassed by conduction. (First published description of an ionization gauge.)	1.33	—	—
Dushman and Found[6] 1921	Molybdenum cylinder 12 millimetres long, 12 millimetres in diameter (collector), with coaxial 3-turn helix of 0.125-millimetre tungsten, 3.65 millimetres inside diameter (anode). Filament, 5 turns 0.125-millimetre tungsten, 2.25 millimetres in diameter; collector carried on separate press.	May be degassed thoroughly by high-frequency or electronic bombardment. Low electrical leakage to collector; linear; shielded from charge variations on glass wall.	3.9 (250 volts)	1.17	Within 2 per cent of unity
Reynolds[7] 1931	As above but cylinder collector 13 millimetres in diameter, 16 millimetres long.	Experimental gauges.	4.4 (250 volts)	1.17	—
	Triode of UV200 construction having a flat plate and hairpin filament	Experimental gauges.	4.2 (125 volts)	1.06	—
Pirani[1]	Telefunken N91A valve as manometer	No details	2.4 (110 volts)	—	—
Jaycox and Weinhart[8] 1931	Plane triode. Oxide filament, grid of 0.010-inch nickel wires; 2 collector plates 1¼ by ¾ inches, ¼ inch from filament.	Readily degassed by electronic bombardment; high collector insulation; shielded; linear; etc.	7.5 (125 volts)	1.7	1.03
	With constructional features to reduce leakage.		13 (200 volts)	—	—
Maloff and Epstein[9] 1938	210-type plane triode; grid positive.	—	2.7 (110 volts)	—	—
Montgomery and Montgomery[10] 1938	247 pentode, grid controlled; screen + plate — collector.	Simple control of electron current.	25‡ (180 volts)	1.17	—
Fogel[4] 1946	Two parallel plates; 1 collector; 1 anode 1⅛ inches by ¾ inch, spaced 9/16 inch apart with a central Vee filament. Mounted on single press but with vapour shield to prevent leakage.	Readily degassed by electronic bombardment or high-frequency; linear; high collector insulation; designed for mass production.	3.75 (200 volts)	1.13	—

* S is measured in microamperes of positive ion current per milliampere of electron current per microbar gas pressure. (1 microbar $=1/1333$ millimetre of mercury.) The tabulated sensitivities refer to argon or air.

† See first paragraph of Section 2.

‡ Because this sensitivity appeared to be high, a copy of the circuit was set up during the present investigation. For the 247 tube tested, $S=2.6$ to 3.5 referred to argon.

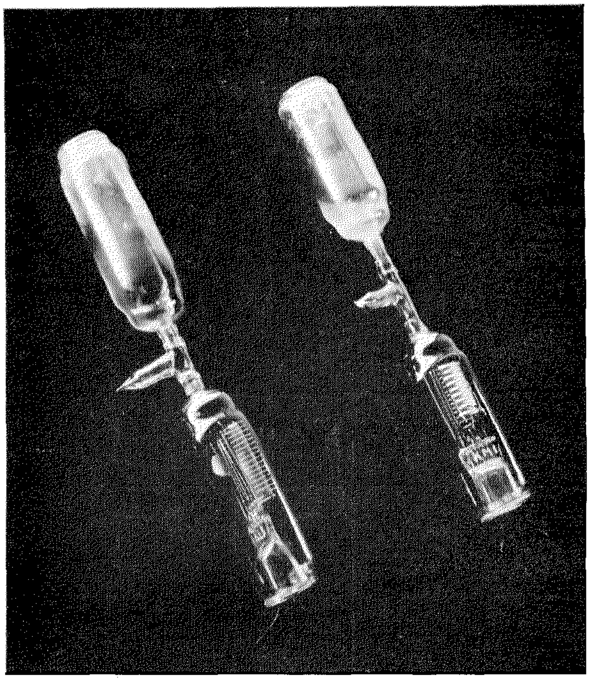

Fig. 4—Sealed-off systems for measurement of manometer characteristics by the mercury-vapour method.

reflector shown in KM2 removes this danger. Tubes of this general design were quite sensitive ($S=18$) and otherwise had good characteristics.

A mercury-vapour method of checking the sensitivities was used in place of the comparison method for most of the tests. Preliminary work showed that reliable results could be obtained from sealed-off tubes containing mercury if double bulbs (Fig. 4) were made up. The second bulb contains a nickel flag enclosing a glass-sealed mercury pill and also a barium getter. After oven bake and normal degassing, the flags are fired and the whole tube again degassed before sealing off. On test, the mercury bulb, neck and part of the manometer are immersed in a thermos of water near zero degrees centigrade. After a short bombardment to clear the electrodes of mercury, the water is stirred and water temperature and ion current readings are taken. Most of the tests were made at a temperature-limited electron current of 200 microamperes but with additional readings at 50, 2000 and 10,000 microamperes to check linearity. The temperature is raised in steps by adding hot water and stirring. Later on, a single large bulb was used for measurements by the same method.

Because the anode current includes the electrons released from ions in addition to the filament emission, for gas pressures at which the ion current is not negligible, it is necessary to choose between constant anode current and constant filament emission. In all tests, the electronic control was set to keep the emission constant. For example, the 71-degree Fahrenheit reading (Table 2) corresponds to an anode current of 210 microamperes. This was considered to be the better alternative because the ejected electrons have a low velocity over most of their path.

The last column of Table 2 gives the mercury sensitivity that must be divided by two for comparison with argon sensitivities quoted earlier.

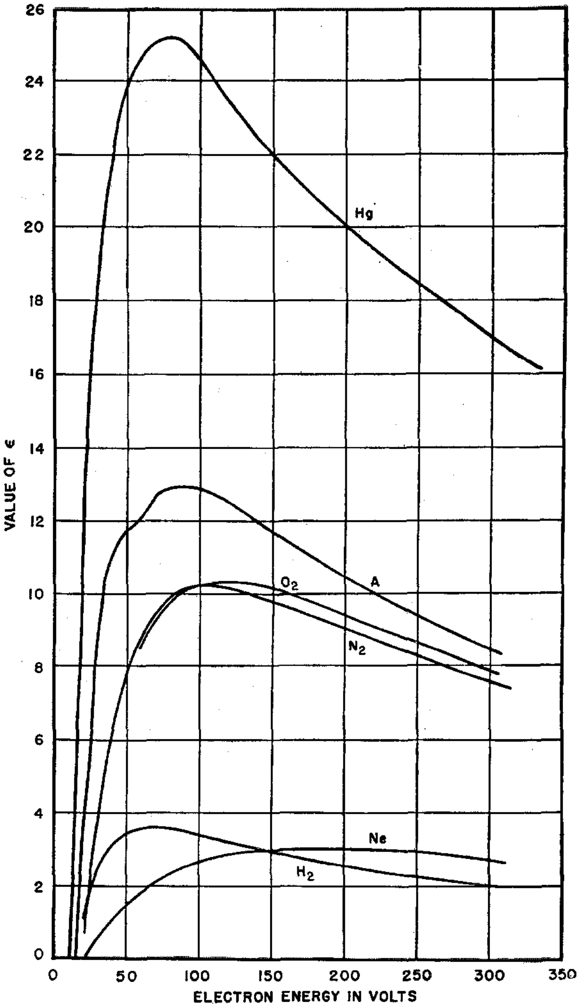

Fig. 5—The efficiency of ionization ε of several gases as a function of electron energy. See Table 2.

2. *Prediction of Electrical Characteristics by Analysis*

Any ionization manometer has the following directly measurable electrical characteristics.

A. Sensitivity S in microamperes per milliampere microbar.

B. Anode characteristic, or variation in S with positive electrode voltage. For simplicity this is specified as A, where

$$A=\frac{S \text{ at } +200 \text{ volts on anode}}{S \text{ at } +150 \text{ volts on anode}}.$$

Because this will vary a little between different gases, all calculated ratios refer to mercury vapour.

C. Negative-collector charateristic. This will be defined similarly as the ratio C, where

$$C=\frac{S \text{ at } -20 \text{ volts}}{S \text{ at } -15 \text{ volts}}.$$

A design objective should be to reduce these ratios as nearly as possible to unity to allow the use of unstabilized voltage supplies. Also, when $C=1$, S will not change if a high resistance is inserted in the collector lead to provide a drive of several volts for an electron-tube microammeter.

The basic information needed in the calculation of the sensitivity of a particular structure is the number of positive charges freed by ionization per electron centimetre of path. This quantity is referred to in the literature as the "efficiency of ionization" as distinct from the "probability of ionization," which indicates the number of ions, whether singly, doubly or multiply charged.[11]

TABLE 2

TEST ON KL3 (EXPERIMENTAL DESIGN)
BY MERCURY-VAPOUR METHOD
Grid Voltage = +125; Collector = −20.

Temperature $T°$ Fahrenheit	Microamperes Ion Current for 200 Microamperes to Grid	Vapour Pressure in Microbars	Apparent Sensitivity S'	Thermal Effusion Correction Factor $\left(\frac{460+71}{460+T}\right)^{\frac{1}{2}}$	True Sensitivity $S' \times$ Factor
32(?)	1.45	0.246	29.4	1.041	30.6
38.5	1.97	0.352	28.0	1.033	28.9
41	2.25	0.40	28.1	1.031	29.0
45	2.80	0.495	28.3	1.027	29.1
48.5	3.30	0.601	27.5	1.023	28.2
50.8	3.72	0.683	27.2	1.021	27.8
55	4.69	0.840	27.9	1.017	28.4
59.5	5.88	1.06	27.8	1.012	28.1
62.8	7.02	1.23	28.5	1.010	28.8
67.6	8.80	1.58	27.9	1.008	28.0
69.5	9.68	1.70	28.5	—	28.5
71	10.40	1.84	28.3	—	28.3

TABLE 3

EFFICIENCIES OF IONIZATION

ϵ = Number of Positive Charges Freed per Centimetre of Electron Path Referred to 1 Millimetre Pressure at 0 Degrees Centigrade

Gas	Observer	Electron Voltage					
		50	100	150	200	250	300
Hg	Jones[17] Bleakney[18] Compton and Van Voorhis[19] Smith[12] Nottingham[28]	23.5 23.6 21.3 17.9 19.0	25.5 24.8 23.1 19.25 14.8	24 21.8 23 18.15 —	22.5 20.4 22.5 16.9 —	20.9 18.5 21.8 15.7 —	19.1 17.1 21 14.6 —
A	Smith[21] Bleakney[23] Compton and Van Voohris[19]	11.68 11.8 10.0	12.9 12.9 12.2	11.83 11.7 12.3	10.53 10.5 11.8	9.43 9.2 11.2	8.58 8.4 10.5
CO	Tate and Smith[24]	9.02	10.9	10.25	9.28	8.4	7.65
N_2	Tate and Smith[24] Compton and Van Voohris[19]	7.88 7.7	10.25 11	9.81 11.7	9.0 11.8	8.26 11.5	7.55 11
He	Smith[21] Compton and Van Voohris[19]	0.86 1.0	1.24 1.7	1.23 1.9	1.15 2	1.06 1.9	0.97 1.9
H_2	Tate and Smith[24] Compton and Van Voohris[19]	3.45 3.6	3.4 4.1	2.94 4.1	2.58 4.0	2.28 3.8	2.03 3.5
Ne	Smith[21] Bleakney[23] Compton and Van Voohris[19]	1.47 1.45 1.1	2.66 2.7 2.5	2.99 3.0 3.3	2.99 3.0 3.5	2.87 2.9 3.6	2.31 2.7 3.8

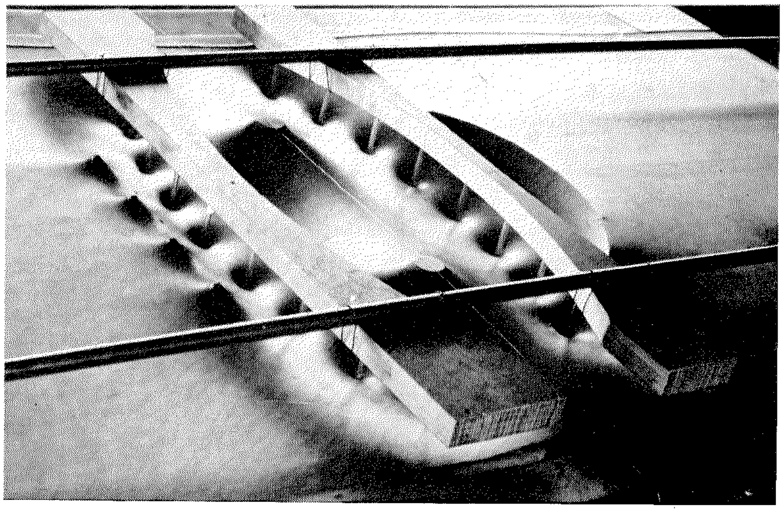

Fig. 6—Rubber-table model of a structure similar to KM2 of Fig. 3. The filaments appear on the left side and the collector on the right.

The efficiency of ionization ϵ may be defined as the number of charges produced per electron centimetre of path in the gas specified, referred to zero degrees centigrade and 1 millimetre pressure. Fig. 5 shows the variation in ϵ as a function of electron voltage for several gases, while Table 3 provides similar information as derived from the literature.

In the method following, the tube to be analysed is split up into a number of sections, perpendicular to the electron flow, chosen so that the voltage gradient may be considered constant over the distance between sections. Space-charge distortion of the static fields is neglected (all the manometers considered use a temperature-limited electron current) and the tubes are assumed to be operating in the low-pressure linear region where only a small proportion of the electrons produce ions.

Consider an electron stream of i_e *milliamperes* moving between parallel equipotential planes l_2 centimetres apart along a normal path.

Let E_1, E_2 = the potentials of these planes measured in volts above the emitting filament.

$G_{1,2}$ = potential gradient, the planes being chosen so that $G_{1,2}$ may be assumed constant.

$G_{1,2}=\dfrac{E_2-E_1}{l_2}$ volts centimetres^{-1}.

i_p = total positive ion current in microamperes produced over this path.

$\phi_{1,2}$ = the fraction of positive charges collected by the negative measuring electrode.

$S_{1,2}$ = the component of sensitivity contributed by the element 1,2.

$=\phi_{1,2}\, i_p/i_e$ when the bulb contains mercury vapour at 1 microbar pressure at 20 degrees centigrade.

Now because the electrons and single ions are of equal but opposite charge,

$$i_p=\frac{1000}{1333}i_e\int_0^{l_2}\epsilon\, dl,$$

whence

$$S_{1,2}=\frac{\phi_{1,2}}{1.333G_{1,2}}\int_{E_1}^{E_2}\epsilon\, dE. \qquad (1)$$

The sensitivity S of the structure is then

$S=S_{0,1}+S_{1,2}+S_{2,3}+\ldots.$

microamperes milliamperes^{-1} microbars^{-1}. (2)

For numerical calculation, since ϵ is not a simply specified function of E, the integrations are performed graphically.

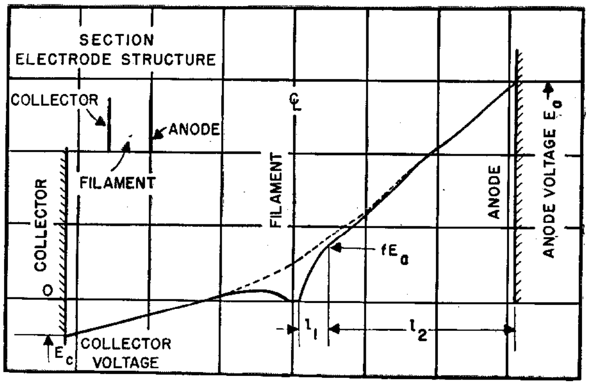

Fig. 7—Potential diagram for a plane plate structure. The dotted line represents the potential between filament strands; the full line indicates the potential in a plane containing the filament.

Evidently, a knowledge of the fields within the electrodes is necessary. These can be calculated from the dimensions only for simple structures, but approximations can be determined by the use of a rubber table model* with the practical advantage that results of changes in the arrangement of electrodes can be judged after some experimentation. A horizontal cross-section of KM2 (Fig. 3) is shown in model form in Fig. 6. Positive electrodes project downwards into the sheet and negative upwards. The grid crosswires are turned vertically in the model following the practice used in calculating triode characteristics by conformal transformation.[14]

2.1 CALCULATION OF S AND A FOR A PLANE PLATE STRUCTURE

The potential diagram and arrangement of electrodes is shown in Fig. 7.

Let $E_a = E_2 =$ potential of the anode plate (volts).

$fE_2 = E_1 =$ potential at distance l_1 centimetres along the electron path to the anode.

The distance l_1 is chosen from the rubber model, or calculated distribution, so that the gradients fE_2/l_1 and $E_2(1-f)/l_2$ may be assumed constant.

From (1),

$$\begin{aligned} 1.33S &= \phi_{0,1}\frac{l_1}{fE_2}\int_0^{fE_2}\epsilon dE + \phi_{1,2}\frac{l_2}{(1-f)E_2}\int_{fE_2}^{E_2}\epsilon dE \\ &= \phi_{0,1}\, l_{1\,(0,\,E_1)} + \phi_{1,2}\, l_2\, Q_{(fE_2,\,E_2)}. \end{aligned} \tag{3}$$

This expression,

$$Q_{(fE_2,\,E_2)} = \frac{1}{(1-f)E_2}\int_{fE_2}^{E_2}\epsilon dE$$

appears in all manometer calculations so that it is worthwhile to provide reference values in a graphical form (Fig. 8). These are calculated from the results of Bleakney[18] and Jones[17] given after Table 4, but referred to room temperature. $Q_{(fE_2,\,E_2)}$ represents simply the charges per

TABLE 4

VALUE OF $0.01\int_0^E \epsilon dE$

ϵ at 0 Degrees Centigrade and 1 Millimetre Pressure

Gas	Observer	Voltage E					
		50	100	150	200	250	300
Hg	Jones[17]	7.0	19.1	31.7	43.6	54.2	64.2
	Bleakney[18]	6.36	18.8	30.3	40.8	50.4	59.3
	Compton and Van Voohris[19]	5.3	16.7	28.3	39.7	50.8	61.4
	Smith[12]	4.74	14.3	23.6	32.4	40.5	48.1
A	Smith[21]	2.8	9.1	15.3	20.9	25.8	30.3
	Bleakney[23]	2.8	9.15	15.3	20.85	25.75	30.15
	Compton and Van Voohris[19]	2.0	7.8	13.9	20	25.8	30.2
CO	Tate and Smith[24]	1.9	7.1	12.4	17.3	21.7	25.7
N_2	Smith[21]	1.4	6.2	11.2	15.9	20.2	24.3
	Compton and Van Voohris[19]	1.3	6.3	12	17.9	23.7	30.5
O_2	Tate and Smith[24]	1.54	6.26	11.4	16.3	20.8	24.9
Ne	Smith[21]	0.22	1.31	2.74	4.24	5.7	6.9
	Bleakney[23]	0.27	1.34	2.78	4.28	5.8	7.2
	Compton and Van Voohris[19]	0.16	1.15	2.56	4.25	6.05	0.9
He	Smith[21]	0.13	0.68	1.31	1.90	2.46	2.97
	Compton and Van Voohris[19]	0.14	0.85	1.78	2.76	3.74	4.74
H_2	Tate and Smith[24]	0.80	2.57	4.14	5.52	6.73	7.79
	Compton and Van Voohris[19]	0.70	2.66	2.74	6.78	8.75	10.5

AVERAGE VALUES OF THE INTEGRAL USED IN ALL CALCULATIONS

20 Degrees Centigrade, 1 Millimetre of Mercury, from References 12, 17 and 18

Volts	30	35	40	50	60	70	80	90	100	125	150	200	250	300
$0.01\int_0^E \epsilon dE$	1.83	2.73	3.72	5.85	8.12	10.5	12.8	15.3	17.6	23.5	29.2	40.0	50.1	59.4

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* Fremlin[13] shows applications of this method in the design of amplifying tubes.

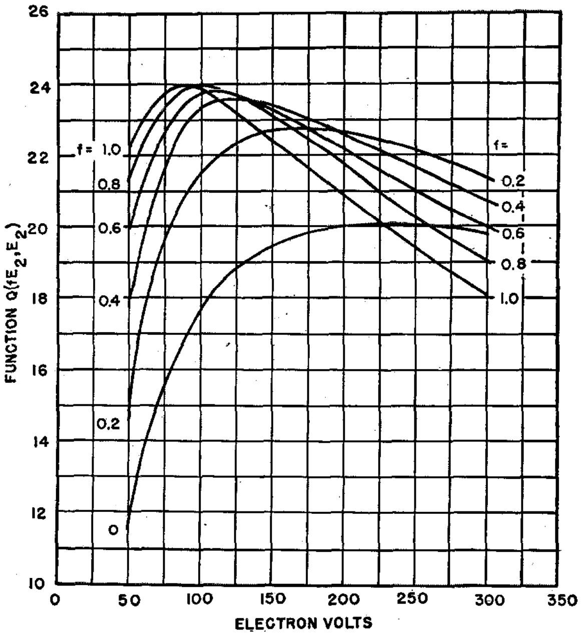

Fig. 8—Reference values for

$$\frac{1}{E_2(1-f)}\int_{fE_2}^{E_2} \epsilon dE = Q(fE_2,\ E_2)$$

for mercury vapour at 20 degrees centigrade.

electron moving between fE_2 and E_2 over a path 1 centimetre long. When $f=1$, the expression becomes $Q=\epsilon$. It will be noted that $Q_{(0,E_1)}$ can be obtained from the curve $f=0$ (Fig. 8). The fractions of ions collected, $\phi_{0,1}$ and $\phi_{1,2}$, are difficult to estimate, but in judging their values it may be noted that in the pressure range considered, before ionization, gas molecules pass freely through the structure rebounding from plates and bulb walls with few collisions between molecules. The average particle energy corresponds to a fraction of an electron volt. Once ionized in the interelectrode space, however, the positive particle is subject to an electrostatic force in a direction opposite to the positive potential gradient. Its final translational energy on reaching the collector will be the collector potential $|E_c|+E$, where E volts above the cathode corresponds to the point at which ionization occurred. Because $|E_c|+E$ will always exceed, say, 20 volts and may average 70 volts or more, and also because a negligibly small proportion of ions recombine except on striking the bulb or collector, the thermal motions have only a small effect, and the ion path could be approximated by ball bearings released from rest at different points along the electron paths on an inverted rubber table model. The bulb walls assume a negative charge under running conditions which means that a proportion could escape around the collector plate of this design.

Fig. 9 compares the measured anode characteristic with that calculated from (3) for $\phi_{0,1}=\phi_{1,2}=1$. The curves suggest that these constants are near $\frac{1}{2}$. Note the improved agreement in slope when the secondary-emission term is included.

2.2 Characteristics of "Long-Path" Design

Fig. 10 is the potential curve for an experimental tube resembling KM2 (Fig. 3). Emitted electrons leave the filament, pass through the sides of the anode, are slowed and returned by the reflector. When the filament plane is reached, those travelling perpendicularly will pass through at low velocity but a proportion deflected to have a velocity component parallel to the filament plane equivalent to the potential "hump" between filament strands, will be reflected and possibly make a second transit of the grid space.

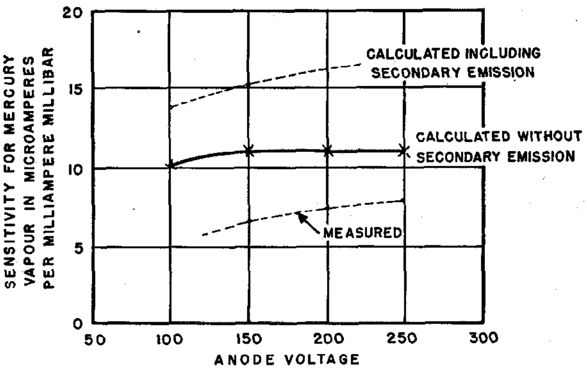

Fig. 9—Calculated and measured anode characteristics of the plane plate structure of Fig. 7. In the calculated curves, ϕ is taken as unity, which is its greatest possible value.

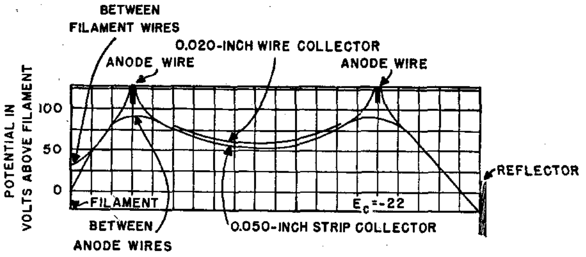

Fig. 10—Potential-distribution curves in a tube of the "long-path" design for two different central collectors. Structures are similar to KM1 and KM2.

By a further approximation, now define the ratio θ as

θ = fraction of the electron stream that passes through one of the grid planes. The fraction $(1-\theta)$ is intercepted.*

Measurements of θ^2 through both planes of a grid similar to that used on KM2 (10.2 turns per inch of 0.008-inch wire) gave $\theta = 0.80$. So that if a second reflector were placed near the filament plane, it would not produce a sufficient gain in sensitivity to justify the complication.

Where l = distance in centimetres between grid planes,
d = distance in centimetres to zero potential plane, just off the reflector,
E_a = applied anode voltage in volts,
E_2 = average potential of electrons passing between grid wires in volts. It is assumed to be equal to the potential midway between the wires.
$E_2/E_a \approx 0.9$, usually, in these tubes,
ϕ_g, ϕ_r = fractions of ion current collected in grid and reflector spaces respectively,

then the expression for the sensitivity is

$$1.33S = \phi_g(\theta+\theta^3)\frac{l}{(1-f)E_2}\int_{fE_2}^{E_2}\epsilon dE + \phi_r 2\theta^2\frac{d}{E_2}\int_0^{E_2}\epsilon dE. \quad (4)$$

Experimental models made with identical constructions but with collectors of decreasing diameter of wire show that ϕ_g remains near unity until 0.02-inch diameter wire is reached. Because the ionizing space is entirely surrounded by a positive electrode, ϕ_g may be set = 1. When d is not too great, also $\phi_r \doteq 1$.

It is an objective to bring A near unity. From an inspection of (4) and Fig. 8, this would seem to be merely a matter of choosing the ratio l/d so that the slopes of the two terms of (4) cancel. But, in fact, secondary emission is an important factor in determining the slope.

2.3 SECONDARY-EMISSION CORRECTION

When electrons (of energy E_a) reach the grid, a small proportion is reflected directly (these are neglected) and others are emitted at all energies between 0 and E_a. A step-by-step procedure is used to estimate the secondary-emission correction, the distribution of energy relation being based on Rudberg's[15] curves, which check roughly with that given by Harries[16] as an average.

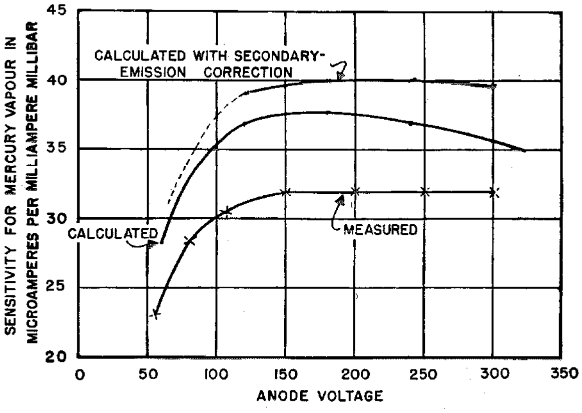

Fig. 11—Calculated and measured anode characteristics of a "long-path" design, KM1. Note that inclusion of the secondary-emission term improves the agreement between the slopes of the characteristics. If the mercury results of Smith[12] are used instead of Jones,[17] the curves agree in magnitude.

If s = secondary emission coefficient at the anode voltage E_a,
k = a fraction of i_e the electron current,
m = a fraction of E_a,

the chosen energy distribution may be represented by the approximation:

Step	1	2	3	4
Current k	0.5	0.22	0.19	0.09
Voltage m	0.1	0.3	0.66	0.93

$k_1+k_2+k_3+k_4=1$ and the secondary coefficient s applying to the incident voltage must be included. From $\theta = 0.8$, the fraction $\theta-\theta^3 = 0.29$ arrives at the back grid and an almost equal current at the front. Consider only the back grid. Now, when $m<(1-f)$ secondary electrons do not pass the potential minimum, so that the first steps add $\Delta_1 S$, $\Delta_2 S$, etc., to the sensitivities where

$$1.33\Delta S \doteq \frac{0.3sk}{(1-\theta^2)}\left[\frac{l}{(1-f)E_2}+\frac{2d}{E_2}\right]\int_0^{mE_a}\epsilon dE \quad (5)$$

$$\doteq \frac{0.3msk}{(1-\theta^2)}\left[\frac{l}{(1-f)}+2d\right]\frac{1}{mE_2}\int_0^{mE_2}\epsilon dE. \quad (6)$$

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* These electrodes resemble grids in construction (see Figs. 3, 6 and 14), but they are always at a positive potential (anodes) in this design.

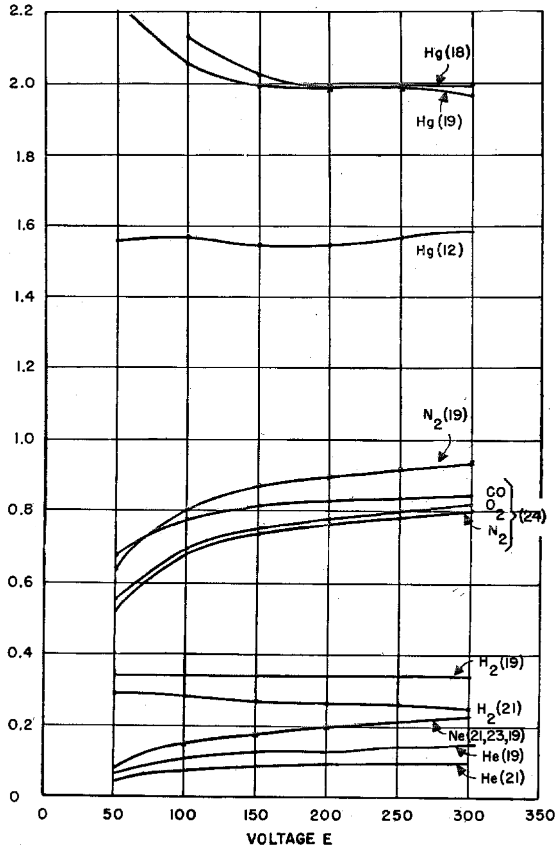

Fig. 12—Probable sensitivity ratios for different gases, plotted from Table 5. Source references are indicated in parentheses.

(The $\frac{1}{1-\theta^2}$ is inserted to sum all the oscillations.) Electrons ejected by the secondaries are neglected. For terms in which $m>(1-f)$, the k component for the fourth step may be inserted in the expression,

$$1.33\Delta_4 S \simeq 0.3ks \times\left[\frac{l\theta}{(1-f)E_2}\int_{(m+f-1)E_2}^{mE_2}\epsilon dE+\frac{2dm}{E_2}\int_0^{mE_2}\epsilon dE\right].$$

Clearly there are approximations in this, only allowable because ΔS is a small correction term.

3. *Sensitivity on Different Gases*

The sensitivity S, defined earlier, varies markedly from gas to gas. A group of ratios of $\frac{S\text{ for gas}}{S\text{ for argon}}$ strictly speaking apply only to one gauge and one anode voltage. The chief gauge characteristics that could cause variation in the ratios are the relative lengths of electron path at high and low energies. For example, if a manometer includes only constant gradients from the anode voltage to zero and secondary emission is neglected, its indications on different gases should correspond with the curves plotted in Fig. 8. Note also that it will have an anode characteristic (on mercury vapour) following curve $f=0$ (Fig. 8), with an additional "rise" (3 per cent between 150 and 200 volts, collector at -15 volts) contributed by the movement of d, the distance to the zero-potential point, outward towards the reflector, if a reflector is used. There is, in fact, a connection between the anode characteristic and the gas ratios for usual designs. Fig. 12 shows that the mercury/argon ratio rises sharply at low voltages.

From this it may be expected that tubes having a greater proportion of low-voltage path than the linear type and/or appreciable secondary emission, would exhibit both a sharply rising anode characteristic and a high mercury/argon ratio. The available evidence supports this although Dushman and Young[25] give a higher ratio (2.73, 2.75 mercury/argon on a cylindrical structure) than can be attributed to this. The converse argument refers more directly to the present design; that is, a tube having an anode characteristic resembling $f=0$ (Fig. 8) should vary in sensitivity between different gases in the ratio of the integrals given in Table 4.

Fig. 12 (from Table 5) shows this ratio as a fraction of anode voltage for several gases. Thus, when the composition of the gas being measured is known, the sensitivity of this design may be predicted from its mercury or argon value by

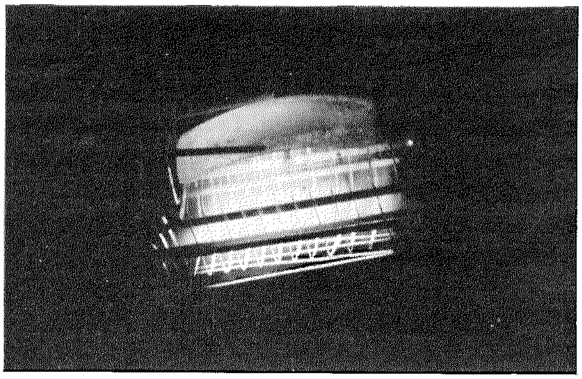

Fig. 13—An electrode system being degassed by electronic bombardment, photographed by its own light. At this temperature, gas is released rapidly.

reference to Fig. 12. A survey of the literature showed that the experimental technique used in the measurement of the efficiency of ionization was improved considerably by Jones[17] and later workers. The results shown in Table 1 were taken from published curves and arranged in an order of preference. Because most of the experimental errors diminish in the calculation of integral ratios, some earlier work is included in Table 5.

4. *Usefulness and Limitations of the Design*

Ionization gauges have been rejected by some workers requiring high accuracy because of the comparatively high rate of gas takeup found in the models used. (See Pirani.[1]) There appear to be several different mechanisms of takeup, which must be considered separately. Apparently, they are:

A. An extremely rapid removal of any gas, found whenever an electrode or the filament reaches a temperature at which its metal has significant vapour pressure. For example, it was found that if the present tube were over-run on bombardment to 28 watts input, a 3-litre tube containing air at 300 microbars was reduced to 0.2 microbar in 30 minutes. A slight dark deposit of nickel appears on the inside of the bulb under these conditions. After metal has been distilled over and the tube switched off, takeup continues for a time but at (in one test) $\frac{1}{6}$th the rate. This method of takeup is an advantage rather than a disadvantage for any application, and applies only to bombardment conditions.

B. There is a slow takeup during operation which depends (probably) on the filament temperature. Takeup of air is found when a tungsten filament is run at near 1300

TABLE 5

VALUE OF $\int_0^E \epsilon_{\text{gas}} dE \Big/ \int_0^E \epsilon_{\text{argon}} dE$

Ratio of Positive Ion Current in the Gas Indicated to the Current in Argon at the Same Pressure for Manometer Tubes Having Approximately Linear Voltage Gradients from E to 0.

Gas	Observer	Voltage E					
		50	100	150	200	250	300
Hg	Bleakney[18]	2.23	2.05	2.0	1.95	1.96	1.97
	Compton and Van Voohris[19]	2.6	2.13	2.03	1.99	1.97	1.97
	Dushman and Young[25]	—	2.73	—	—	—	—
	Smith[12]	1.7	1.6	1.54	1.55	1.57	1.58
A	All Observers	1	1	1	1	1	1
CO	Tate and Smith[24]	0.7	0.78	0.81	0.83	0.84	0.85
	Dushman and Found[6]	—	0.81	—	—	—	—
N_2	Smith[21]	0.52	0.68	0.73	0.76	0.78	0.80
	Compton and Van Voohris[19]	0.64	0.80	0.87	0.87	0.92	0.94
	Hughes and Klein[22]	—	0.9	0.95	0.97	0.97	0.98
	Dushman and Found[6]	—	0.88	—	—	—	—
	Dushman and Found[25]	—	0.89	—	—	—	—
O_2	Tate and Smith[24]	0.56	0.69	0.74	0.78	0.80	0.82
Ne	Smith[21]	0.08	0.14	0.18	0.20	0.22	0.23
	Bleakney[23]	0.09	0.146	0.18	0.20	0.224	0.24
	Compton and Van Voohris[19]	0.08	0.15	0.18	0.21	0.23	0.25
	Hughes and Klein[22]	0.07	0.14	0.185	0.21	—	0.242
	Dushman and Young[25]	—	0.196	—	—	—	
He	Smith[21]	0.045	0.075	0.085	0.091	0.095	0.097
	Compton and Van Voohris[19]	0.07	0.11	0.13	0.14	0.145	0.15
	Hughes and Klein[22]	0.06	0.09	0.10	0.11	—	0.12
	Dushman and Young[25]	—	0.14	—	—	—	—
H_2	Smith[21]	0.29	0.28	0.27	0.264	0.261	0.257
	Compton and Van Voohris[19]	0.34	0.34	0.34	0.339	0.339	0.338
	Hughes and Klein[22]	0.19	0.28	0.30	0.30	—	0.30
	Dushman and Young[25]	—	0.39	—	—	—	—

Observations from references 6 and 25 are from experimental measurements on manometers. Strictly speaking, they are not comparable and are included only for general comparison. Nonlinearity, secondary emission and the other factors discussed earlier may account for the difference, although the argon measurement of reference 25 appears to include most of the discrepancy.

degrees centigrade. An oxide-coated-filament tube showed rates of

17 milliamperes—0.6 cubic centimetres per second
8.6 milliamperes—0.3 cubic centimetres per second
0.4 milliamperes—0.1–0.2 cubic centimetres per second

(pressure approximately 0.1 microbar) for an almost worn-out manometer having a fairly bright filament; a new tube would probably be lower.

C. Blears[27] notes that in the measurement of an organic vapour, a 0.7-litre-per-second tubulation gave approximately 1/10th the pressure reading of an 83-litre-per-second connection. It is thought that the filament temperature would be important in determining this rate and similar tests may be made later with the present oxide filament, which runs "black" when the electron current is reduced to 200 microamperes.

5. *Gas Heating*

For precise measurements, a small correction must be made for gas heating.[26] For tubes running at low electron currents, the openness of the structure (Fig. 14) suggests that the bulb wall temperature may be taken as the gas temperature with little error.

6. *Insulation and Photoelectric Currents*

Simplicity of manufacture required that all elements should mount from the press. Two precautions have been taken to reduce electrical leakage:

A. The collector leads pass through glass sleeves projecting internally and externally.

B. The high potential anode is separated by the (low-potential) filament leads from the collector so that a slight "guard ring" protection is obtained.

Insulation resistance measurements made between collector and all other electrodes for tubes after overrunning on bombardment (darkened bulbs) give values of 2×10^{10} ohms and for new tubes, 5×10^{10} ohms. Photoelectric currents are greater than leakage currents in bright daylight, and for most precise work the bulb should be screened from light and the filament kept cool by running at a low electron current.

7. *Sensitivity*

"Long-path" designs similar to KM2 have good electrical characteristics. But a drawback is the close spacing needed between the collector loop inside the grid and the grid sidewires, to keep the fraction f above 0.6. This complicates the assembly operation and makes the structure sensitive to bumps and vibration.

A satisfactory solution is the replacement of the central collector frame by a "virtual collector." This consists of a pair of openings cut in the back surface of the grid as shown in Fig. 14. Ions are accelerated towards these openings from almost all points in the grid space. With this construction, f may be set at any value by the adjustment of the sizes of the grid openings. It is then comparatively easy to set f and d so that the anode characteristic A is unity. Also, the structure has only 3 simple electrodes: filament, grid and reflector (collector), with no critical spacings.

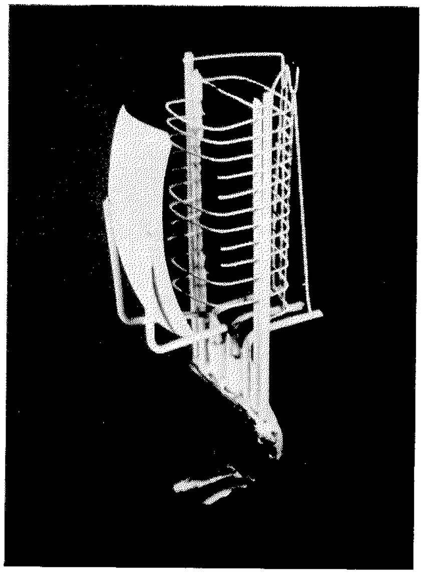

Fig. 14—The electrode system of the simplified design. The central frame seen in KM2 of Figs. 3 and 6 has been replaced by a pair of openings in the grid.

For the dimensions tentatively adopted, tests on experimental tubes show

S (argon)
$=15$ microamperes per milliampere microbar

for voltages between 125 and 200. For other gases, the curves shown in Fig. 12 may be used to increase or decrease this figure. The value of A varies with the type of gas, but for mercury or air in the linear range, A is within 3 per cent of unity and C is within 1 per cent of unity.

8. Linearity at High Gas Pressures

Tests were made to find the high-pressure limit of linear operation on several experimental designs. These confirmed Dushman's tests, showing that the high pressure range is extended considerably by the use of low anode currents. At 200 microamperes electron current, a high limit was found for mercury vapour at 1.8 microbars. At this pressure, when the anode voltage was increased above 180, the sensitivity dropped suddenly to nearly one half. On reducing the anode voltage, the low sensitivity was maintained to 120 volts when it rose exactly to its former value. The cycle could be repeated accurately and was caused by comparatively highly conductive paths forming between filament and anode. In the linear high-sensitivity region, the glow is dull and uniform.

9. Conclusion

It is hoped that the gauge may find application in fields of research and manufacture.

10. Acknowledgments

I wish to thank Mr. K. S. Brown, Valve Department Manager at Standard Telephones and Cables Pty. Limited for encouragement to complete and record this investigation. Notably, Mr. J. Howes has carried out much work on the project and many other members of the department have assisted with their particular skills.

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Long-Distance Telephone Communication Circuits*

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WHILST THE voice-frequency-circuit mileage in use compares favorably with that of the newer carrier systems, attention in engineering and installation is being focused on multi-channel carrier telephony. A brief survey of the characteristics and fields of use of carrier telephony on open wire, cable, and coaxial conductors is presented. By standardising on a basic group of 12 channels and appropriate frequency bands for all types of carrier systems, a considerable amount of interworking flexibility is obtained and manufacturing costs are reduced through common use of filters and other components.

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1. *Voice-Frequency Circuits*

Although the development of carrier circuits in recent years has rather stolen the limelight and has tended to relegate voice-frequency circuits into the background as far as the communications engineer is concerned, this is largely a technical eclipse, and it must not be forgotten that the amount of audio-frequency circuit mileage still compares with that provided by carrier means, and is therefore of significant importance to the traffic and financial side of an administration or operating company. As technical advances are made, the dominant long-distance-communication system of one day is replaced in importance by another. But the general growth of telecommunication requirements seems to ensure that in well-planned and well-equipped networks, the older systems always retain some uses, generally in a somewhat different role.

However, the technical and operating aspects of audio-frequency circuits are so well known that it is not proposed to deal with them in any detail except to say, in passing, that a case for audio-frequency cables still exists, particularly on very short routes, and that the presence of the metallic conductors for each circuit may enable simple and inexpensive signalling facilities to be obtained.

2. *Carrier Circuits*

2.1 GENERAL

Land-line telephone transmission by carrier means has so far followed the technique of multiplexing on a frequency basis; that is to say, the audio-frequency band has been transposed to occupy a position in another part of the frequency range, and is then retransposed to the audio range at the receiving end. Other channels are obtained by transposing to adjacent parts of the frequency spectrum, permitting several channel frequencies to be transmitted over a common line and through common amplifiers, thus saving line plant.

Since attenuation increases with frequency, it is the aim of the engineer to use the frequency range available on the line to the best advantage, and this means limitation of the transmitted band to that considered adequate and desirable, and locating such bands as close to each other as possible without causing undue interference between channels. Limitation of the bandwidth and the segregation of channels is performed by filters.

It is obvious that the best system from both technical and economic standpoints must be a compromise between many factors. Thus, spacing channels further apart eases the filter requirements and therefore the costs, but it also means that fewer channels can be transmitted within a given frequency range. To increase the frequency range, a greater number of amplifiers are needed to give a closer spacing of repeaters. Similarly, compromises in amplifier output levels must be achieved, and a balance struck between high outputs with large valves and consequently expensive power supplies, or smaller outputs and valves, but closer repeater spacing.

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* From a paper presented before the Associazione Elettrotecnia Italiana in Rome, Turin, and Milan during July, 1947.

These are just one or two of the factors that readily come to mind, but a great deal of thought has been given to such problems. To refer to only two instances of such studies, I might mention Jacobsen's paper,[1] which shows that the power-handling capacity of an amplifier carrying many speech channels can be determined from knowledge of the harmonic levels produced by the amplifier when loaded with a sine wave of known power. Also of interest is a paper by Hodgson and Roche[2] on the broad subject of the planning of systems, either individually or as links in long circuits, with special reference to the repeater spacing and output power and their influence on overall circuit noise.

Apart from these and other fundamental problems, there are details of circuit design with their bearing on manufacturing technique that must be solved effectively if the most economical equipment is to be obtained; for example, the number of components required and the space occupied can be reduced by combining several functions in one circuit. The incorporation of channel modulators designed to act as volume limiters is one illustration of this point. Another is the employment of the channel demodulator amplifier input coil to form part of the low-pass filter and equaliser, an economy possible in the 60–108-kilocycle crystal-filter system because the unwanted frequencies are well removed from the wanted band.

Of course, in considering technical products the cheapest in first cost is not necessarily the most economical in the long run, and perhaps nowhere is this more true than in the communication field, where ease of maintenance, stability, and provision for ordered growth are all of the greatest importance.

To obtain such ordered growth, it is obviously advantageous to build in units, and the basic unit, which the International Standard Electric Corporation uses (and which conforms to the recommendations of the Comité Consultatif International Téléphonique), is the group of 12 channels. The number 12 does not in itself have any predominant significance; indeed, 10 might find most favour if other factors were equal, but taken in conjunction with the frequency spacing of the channels and the types of filters used, this number is particularly suitable and contributes to ease of manufacture and therefore to economy.

2.2 12-Channel Carrier-on-Cable System

A block schematic of the 12-channel carrier-on-cable system is shown in Fig. 1. Since this system is widely known, it will not be discussed in detail here.

Very briefly, however, the voice-frequency currents pass from the 2-wire line through the terminating equipment to the channel modulator, after which the lower sideband is selected by the band-pass filter. The carrier frequencies supplied to the 12 modulators range from 64 to 108 kilocycles in steps of 4 kilocycles, so that when the output currents from the 12 filters are united in the common transmitting path, they comprise a series of nominal 4-kilocycle bands lying between 60 and 108 kilocycles. The carrier currents themselves are substantially suppressed by the filters. This basic group-frequency band is then modulated at 120 kilocycles and the lower sideband, occupying the range of 12–60 kilocycles, is transmitted to the line through the transmitting amplifiers.

The receiving circuit reverses this process, the incoming group at 12–60 kilocycles being transposed to the 60–108-kilocycle range from which the individual channel band-pass filters select the appropriate bands for the associated channel demodulators, voice-frequency amplifiers, and terminations.

It will be noticed that all of the portion at the left of the diagram comprises the frequency-generating equipment; as this supplies many systems, its contribution to the sum total of apparatus in a quantity sense is unduly accentuated by such a diagram. This is, of course, the heart of the system, and so must be made to be completely reliable, even if this involves what may at first seem to be quite a heavy cost. However, when this cost is divided among the individual channels, it does not represent a high proportion of the cost per channel.

When apportioning the amount and cost of the different classes of apparatus, it should be noted

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[2] K. G. Hodgson and A. H. Roche, "Design Factors Influencing the Economical Size and Spacing of Multi-Channel Telephone Repeaters," *Electrical Communication*, v. 19, pp. 100–107; October, 1940.

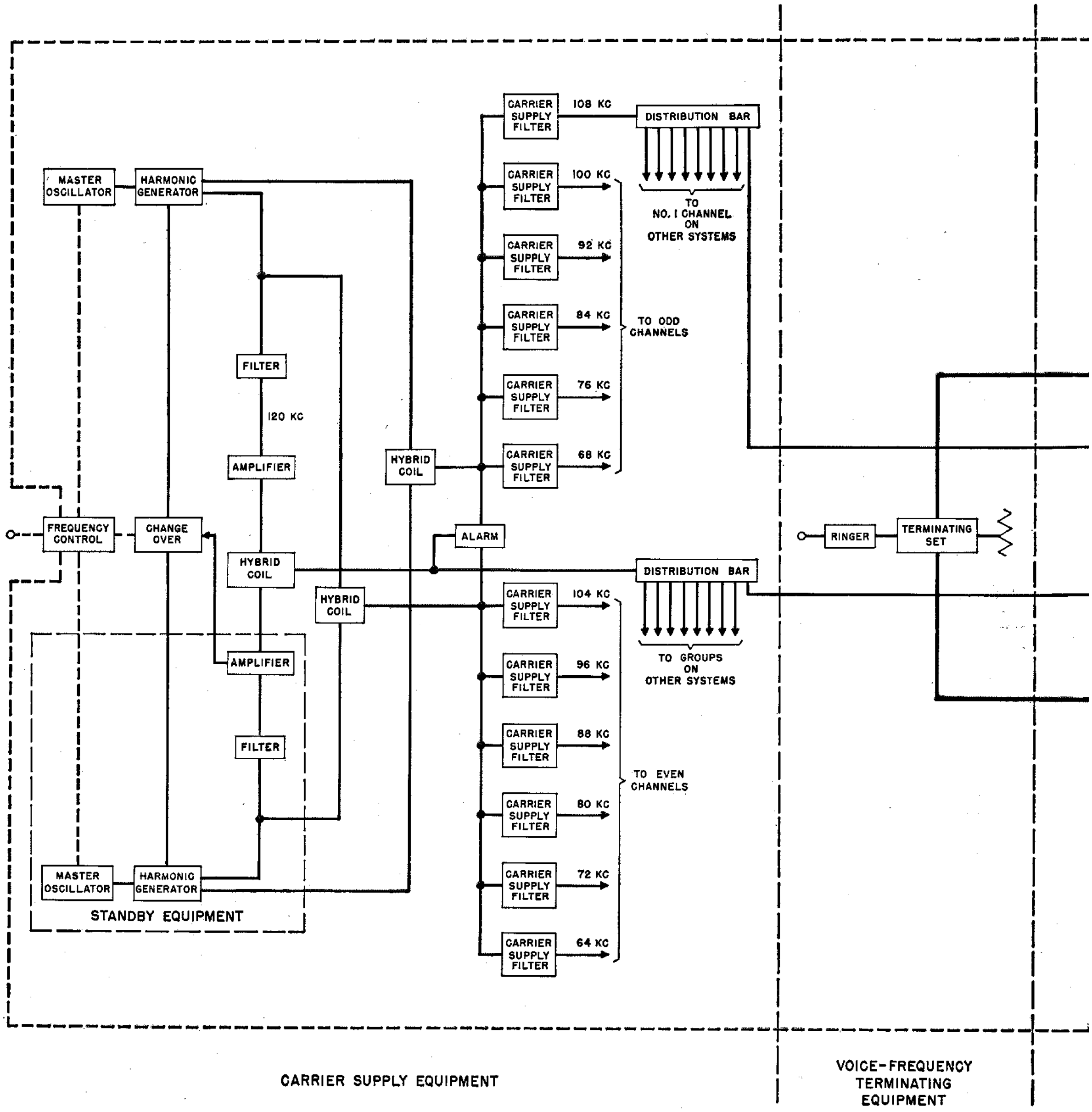
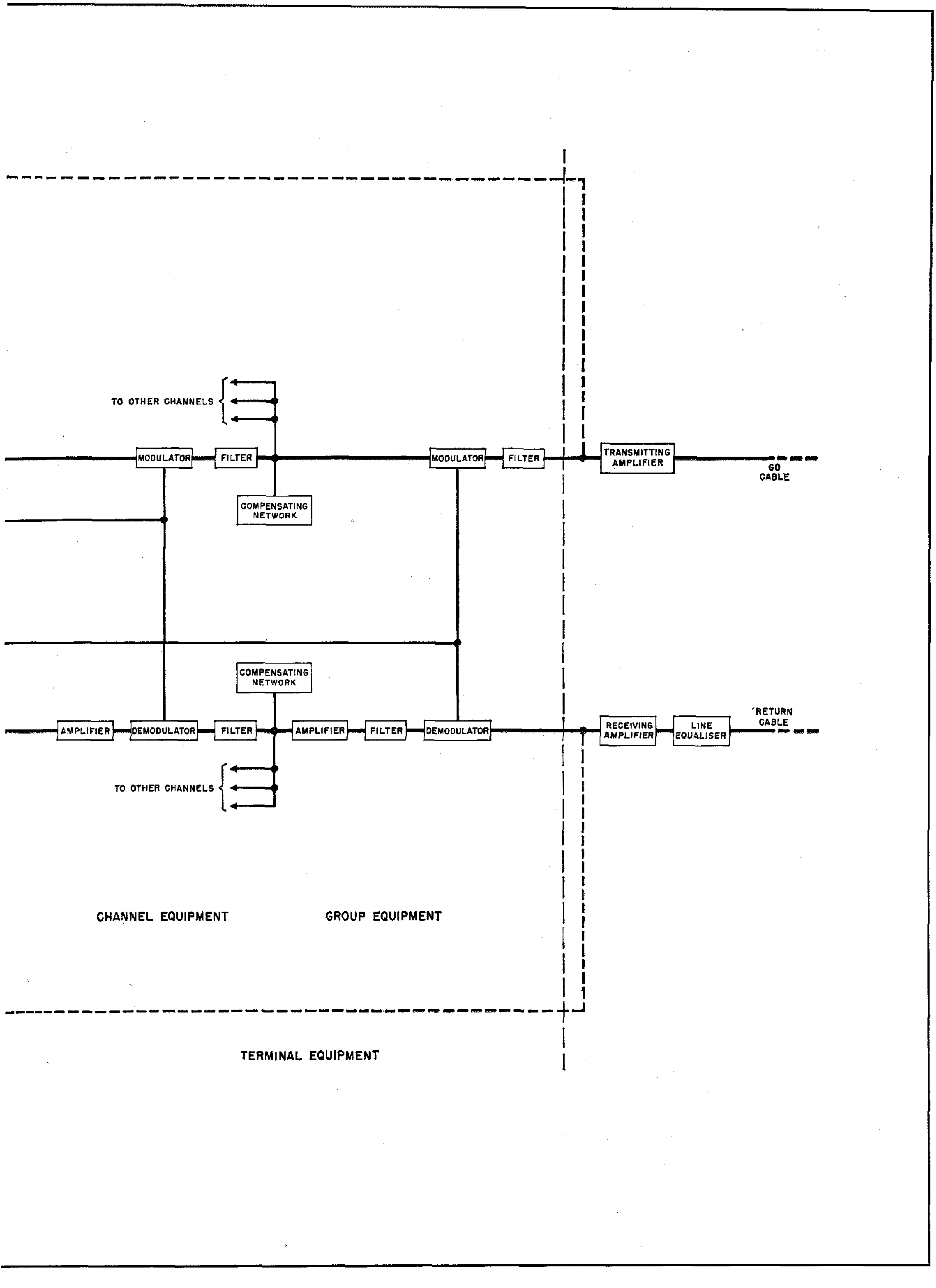


Fig. 1—Block diagram of 12-channel carrier-on-cable terminal equipment.


TO OTHER CHANNELS
MODULATOR
FILTER
MODULATOR
FILTER
TRANSMITTING AMPLIFIER
GO CABLE
COMPENSATING NETWORK
COMPENSATING NETWORK
AMPLIFIER
DEMODULATOR
FILTER
AMPLIFIER
FILTER
DEMODULATOR
RECEIVING AMPLIFIER
LINE EQUALISER
RETURN CABLE
TO OTHER CHANNELS
CHANNEL EQUIPMENT
GROUP EQUIPMENT
TERMINAL EQUIPMENT

that a signalling unit and terminating equipment are required for each channel. Moreover, the terminal equipment proper comprises a modulator and band-pass filter per channel in the sending path, and similar equipment, together with a small single-valve amplifier, in the receiving path for each channel.

Equipment common to all 12 channels is made up of the filter-compensating networks, the group modulator, demodulator, filters, receiving amplifier, line-transmitting amplifier, and receiving equaliser and amplifier.

It is important to realise that whereas the carrier equipment as shown can supply 120 individual channel terminals (or 480 channel terminals with the addition of 12 carrier amplifiers), and that the group and line equipment serves 12 channel terminals, each individual channel terminal requires 2 band-pass filters and one signalling unit as well as certain smaller items. The cost of the overall terminal equipment lies primarily, therefore, in the channel filters and the signalling unit.

It would be possible to achieve a similar technical result without crystal filters by using other modulation processes, but experience shows, after repeated checks at intervals, that the crystal-filter method proves more economical in production and use.

The 12-channel cable system is designed to work on a 4-wire basis using one cable for transmission in one direction and another for transmission in the opposite direction. This results in a very simple repeater comprising an equaliser and amplifier in each direction.

Repeater spacing is normally based on a loss of 6 nepers (52 decibels) at 60 kilocycles, which means that repeaters will be about 32 kilometers (19.9 miles) apart, although some variation may be allowed.

2.3 24-CHANNEL CARRIER-ON-CABLE SYSTEM

With closer repeater spacing, more accurate cable balancing, and suitable line equalisers, another group of 12 channels can be operated on the same cable pairs, thus doubling the circuit capacity. This can be achieved in a simple manner by adding a basic group at 60–108 kilocycles to the nominal 12-channel group at 12–60 kilocycles. Fig. 2 shows the 24-channel system built up from the 12-channel system. High-pass and low-pass crystal-filter combinations are used to separate and combine the two groups, and additional amplifiers are employed to compensate for the loss in these filters and to produce the required levels at the flexibility points. These flexibility points are used for switching channel groups and high-frequency lines as required, and are termed "Group-Distribution Frame" and "High-Frequency-Repeater Distribution Frame." At the group-distribution frame, channels are at the basic group frequency of 60–108 kilocycles. The sending level at this point is −0.92 neper (−8 decibels), and the receiving level is −4.25 nepers (−37 decibels) with reference to the transmitting toll-test-board level.

At the high-frequency-repeater distribution frame, the frequencies lie between 12 and 108 kilocycles with a normal level of +0.575 neper (5 decibels) for both sending and receiving.

The line equalisers and amplifiers are designed to cover the range 12–108 kilocycles, and, in fact, the 24-channel amplifier is now standard equipment whether the circuits are to be used for 12- or 24-channel operation. The repeater spacing for 24-channel systems is about 23–25 kilometers (14.3–15.5 miles) for 1-millimeter (18 American Wire Gauge) conductors, or 29–32 kilometers (18.0–19.9 miles) for 1.3-millimeter (16 American Wire Gauge) pairs. Where circuit groups and distances do not justify the use of coaxial systems, and operation of carrier on multi-pair cables is economical, the 24-channel system, in the author's opinion, represents the most efficient and economical method of using the cables. This view is confirmed by considered opinion in the British Post Office, as expressed by Mr. Chamney in his recent paper.

2.4 36- AND 48-CHANNEL CARRIER-ON-CABLE SYSTEMS

A great deal of study has been given to systems employing more than 24 channels, and 36- and 48-channel systems and suitable repeaters can be supplied, but there are two main points that require careful consideration.

In the first place, they are generally uneconomical when compared with 24-channel systems, and certainly when compared with systems using coaxial lines.

Secondly, experience shows that extension beyond 24 channels introduces difficult far-end-crosstalk problems such that the higher-frequency channels may be restricted to comparatively short distances and rendered unsuitable for long national or international circuits. In certain special cases, such systems may possibly be justified, and the International Standard Electric Corporation has designed and produced 48-channel equipment for the Dutch Administration. The distances involved in Holland are abnormally small, however, and existing repeater stations are available at extremely short intervals.

2.5 COAXIAL SYSTEM

It should be noted that all the systems so far mentioned are built up from the basic group of 12 channels, transposed or not, as the requirements of the system demand.

This principle is followed in the coaxial system, in which the *basic group* of 12 channels is used as a unit to build a *basic supergroup* consisting of five such groups, i.e., 60 channels. Ten basic supergroups are then transposed to adjacent positions in the frequency spectrum and transmitted over a coaxial tube consisting of a central conductor positioned by insulating washers inside an outer copper cylinder, which serves as the other conductor. One such tube is used for transmission in one direction and another tube in the same cable is used for transmission in the reverse direction.

Fig. 3 shows the frequency transpositions involved. Five basic groups in the range of 60–108 kilocycles are modulated by carriers of 420, 468, 516, 564 and 612 kilocycles, and the lower sidebands are then selected by filters to produce the basic supergroup of 60 channels in the range 312–552 kilocycles. Nine such basic supergroups are modulated by carrier frequencies of 612, 1116, 1364, up to 2604 and 2852 kilocycles. The lower sidebands of these, together with an untransposed basic supergroup, produce a series of 600 channel bands for transmission on the line in the range 60–2540 kilocycles. The untransposed basic supergroup forms supergroup 2 on the line.

2.5.1 Coaxial Terminal Equipment

The arrangement of apparatus at terminals, excluding the carrier-supply equipment and terminal repeaters, is shown in Fig. 4. Normal channel bay equipment, similar to that required for systems described above, is used. The group-frequency-transposing equipment modulates five groups to produce a basic supergroup, and it will be noticed that the groups are combined by means of a hybrid coil with alternate group-frequency band-pass filters connected in parallel. This eases the filter requirements, since impedance irregularities at the edges of the filters due to proximity are avoided, and the effective discrimination is improved. The individual basic supergroups are amplified, modulated, and selected by filters, and are then combined by means of a hybrid coil before passing through a common amplifier to the transmitting line-repeater equipment. A total of 480 channels (8 supergroups) are handled in this way, while two more supergroups may be added in the manner shown.

The receiving portion of the terminal circuit is similar, except that the process is reversed. The extreme economy in amplifiers and valve circuits on this system—combined with adequate provision for reliability—should be noted. In the whole of the sending path there is one single-stage amplifier per supergroup, apart from the auxiliary transmitting amplifiers of which there are but 1 for 8 supergroups or 2 for 10 supergroups. In the receiving path, there is a single-stage supergroup amplifier, a 2-stage group amplifier, and a single-stage channel amplifier. Because the supergroup amplifiers are shared by 60 channels, the valves are duplicated, so that failure of one valve will not cause a large number of circuits to fail. The auxiliary transmitting amplifiers are similar, i.e., single-stage but containing two valves.

Summarising, 600 individual channel terminals require 600 single-stage channel amplifiers, 50 2-stage group amplifiers, 20 single-stage supergroup amplifiers (with duplicate valves), and 2 single-stage auxiliary transmitting amplifiers (with duplicate valves).

When the carrier-supply and terminal-repeater equipment is also added, the number of valves per individual channel terminal at a fully equipped 10-supergroup terminal, excluding signalling equipment, is less than 1.4. This is a considerable achievement, and is reflected in ease of maintenance, low power requirements, and reduced annual charges.

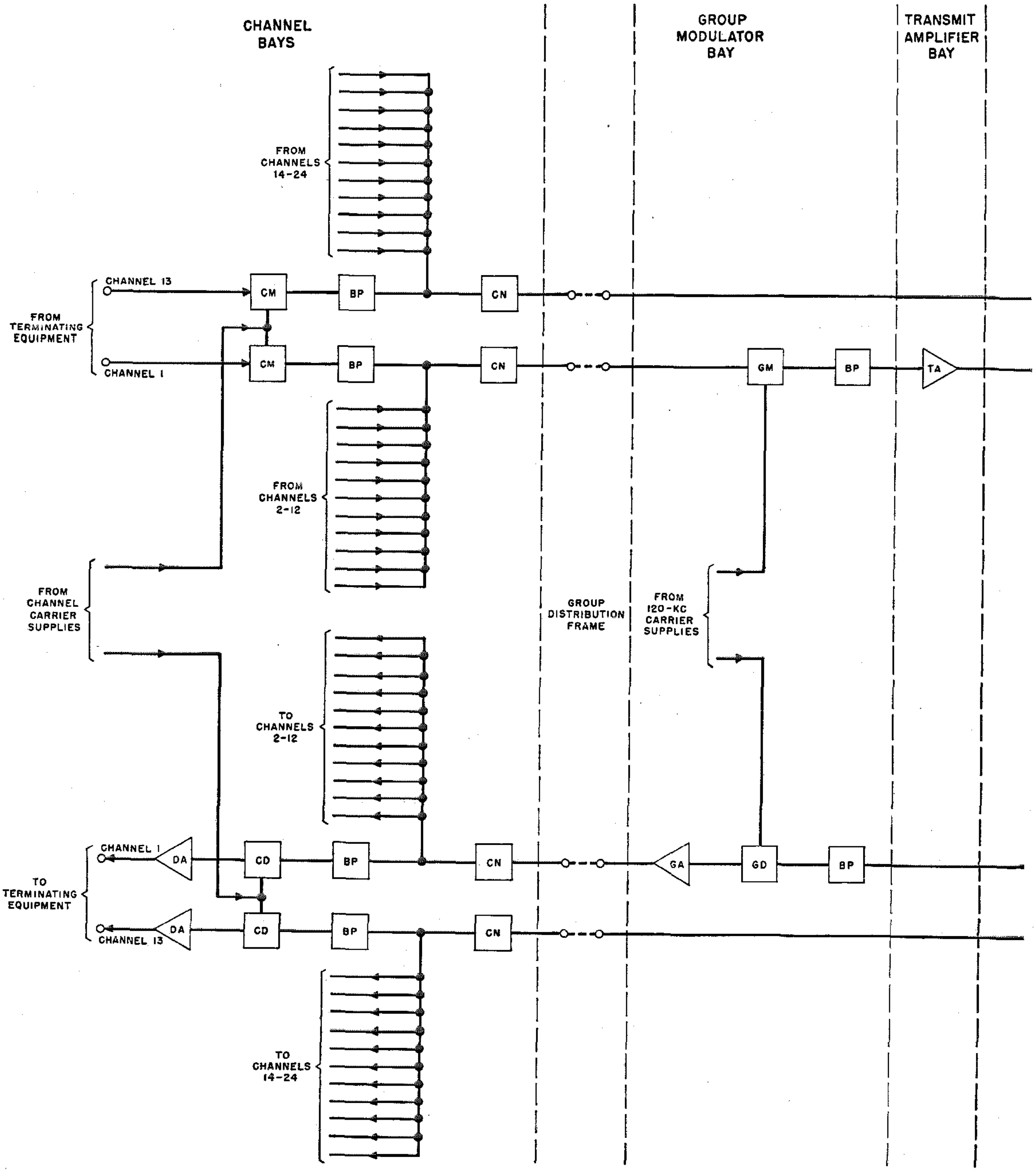
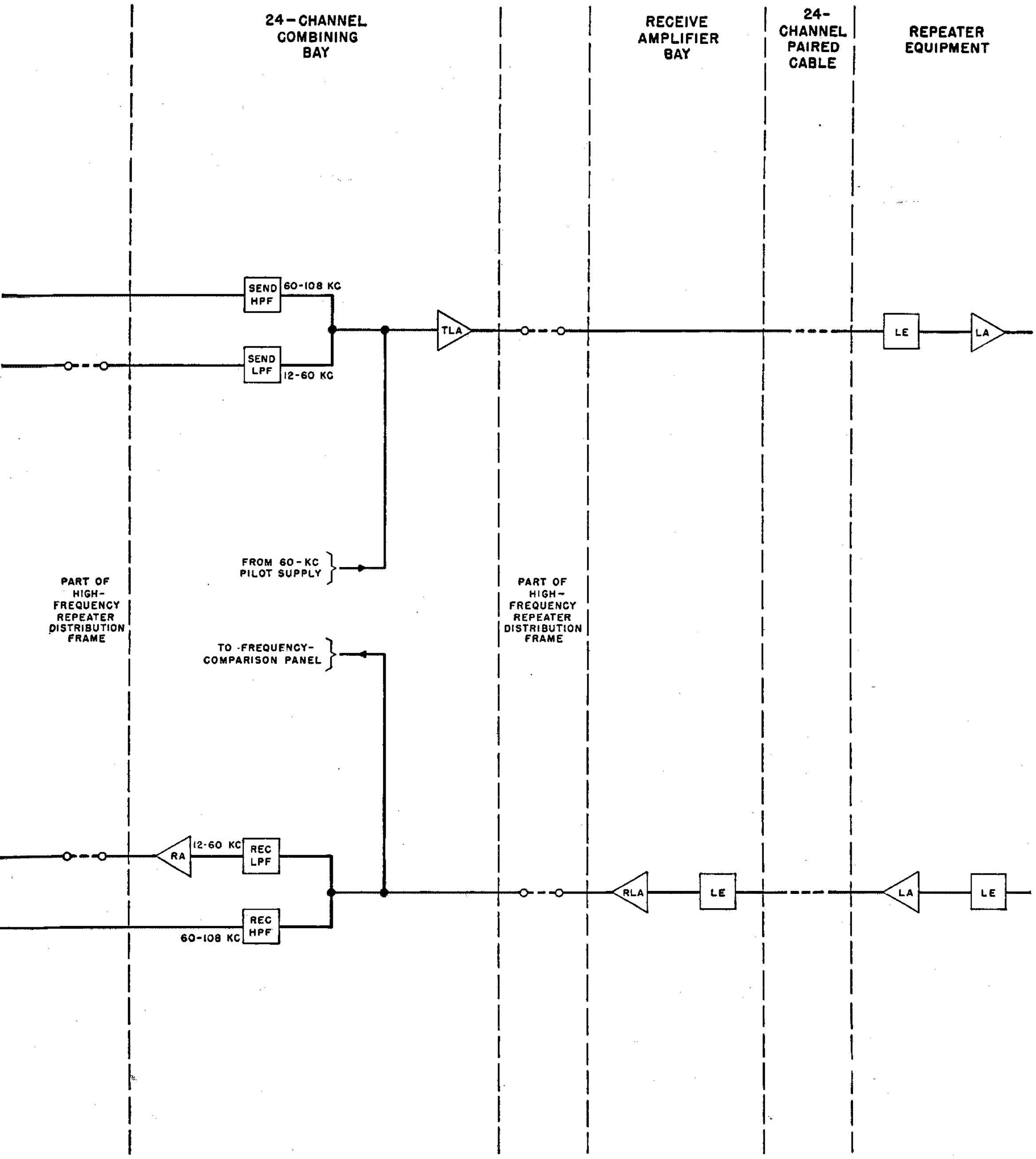


Fig. 2—Block diagram of transmission circuits for terminal and repeater. The letters in the boxes indicate the following: BP, Band-pass filter. CD, Channel demodulator. CM, Channel modulator. CN, Compensating network. DA, Demodulator amplifier. GA, Group amplifier. GD, Group demodulator. GM, group modu-



lator. HPF, High-pass filter. LA, Line amplifier. LE, Line equalizer. LPF, Low-pass filter. RA, Receive amplifier. RF, Receive filter. RLA, Receive line amplifier. TA, Transmit amplifier. TLA, Transmit line amplifier.

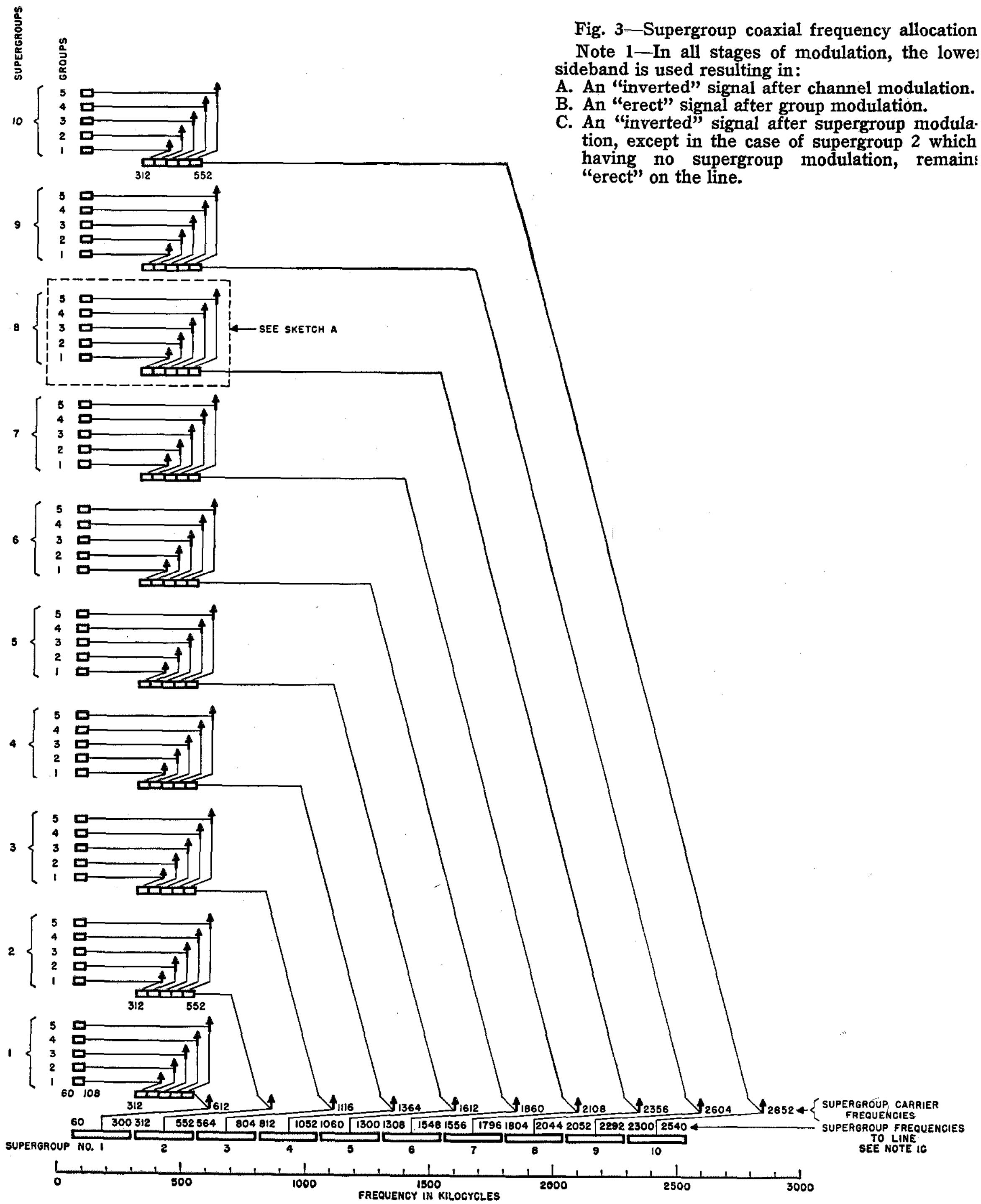
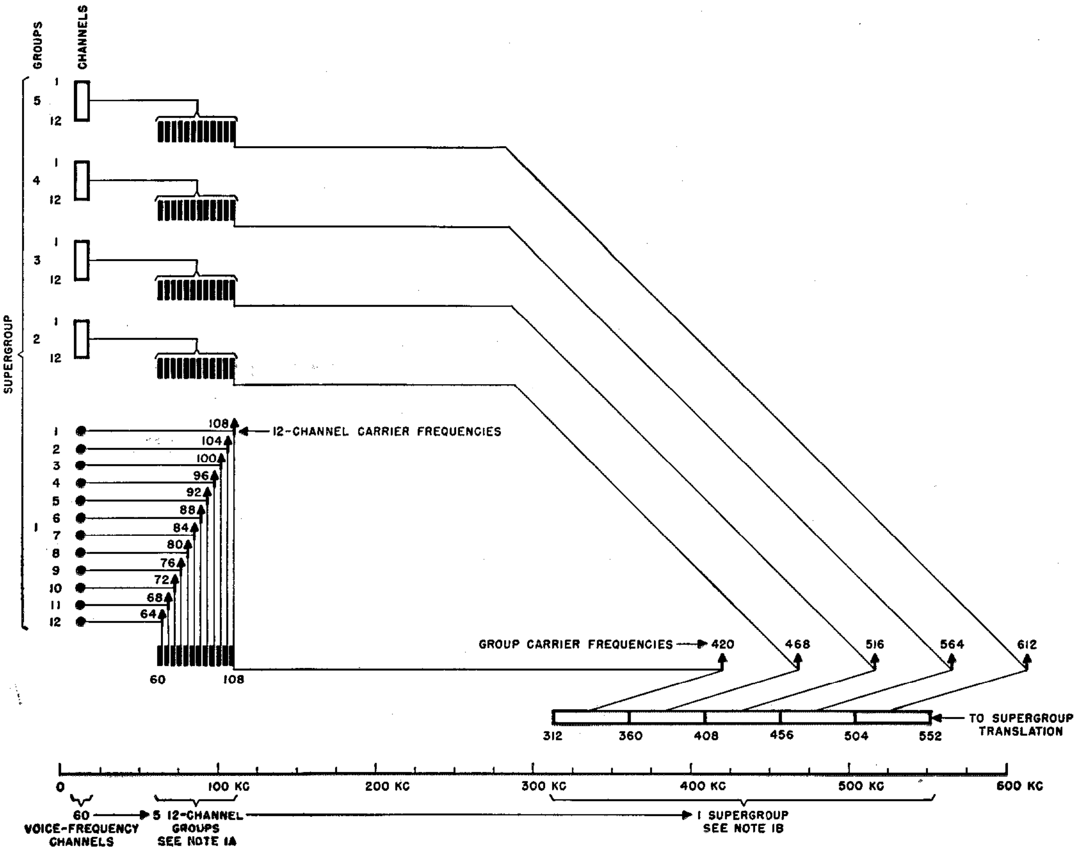


Fig. 3—Supergroup coaxial frequency allocation

Note 1—In all stages of modulation, the lowe sideband is used resulting in:

A. An "inverted" signal after channel modulation.
B. An "erect" signal after group modulation.
C. An "inverted" signal after supergroup modulation, except in the case of supergroup 2 which having no supergroup modulation, remain "erect" on the line.



SKETCH A
TYPICAL SUPERGROUP PRIOR
TO SUPERGROUP TRANSLATION

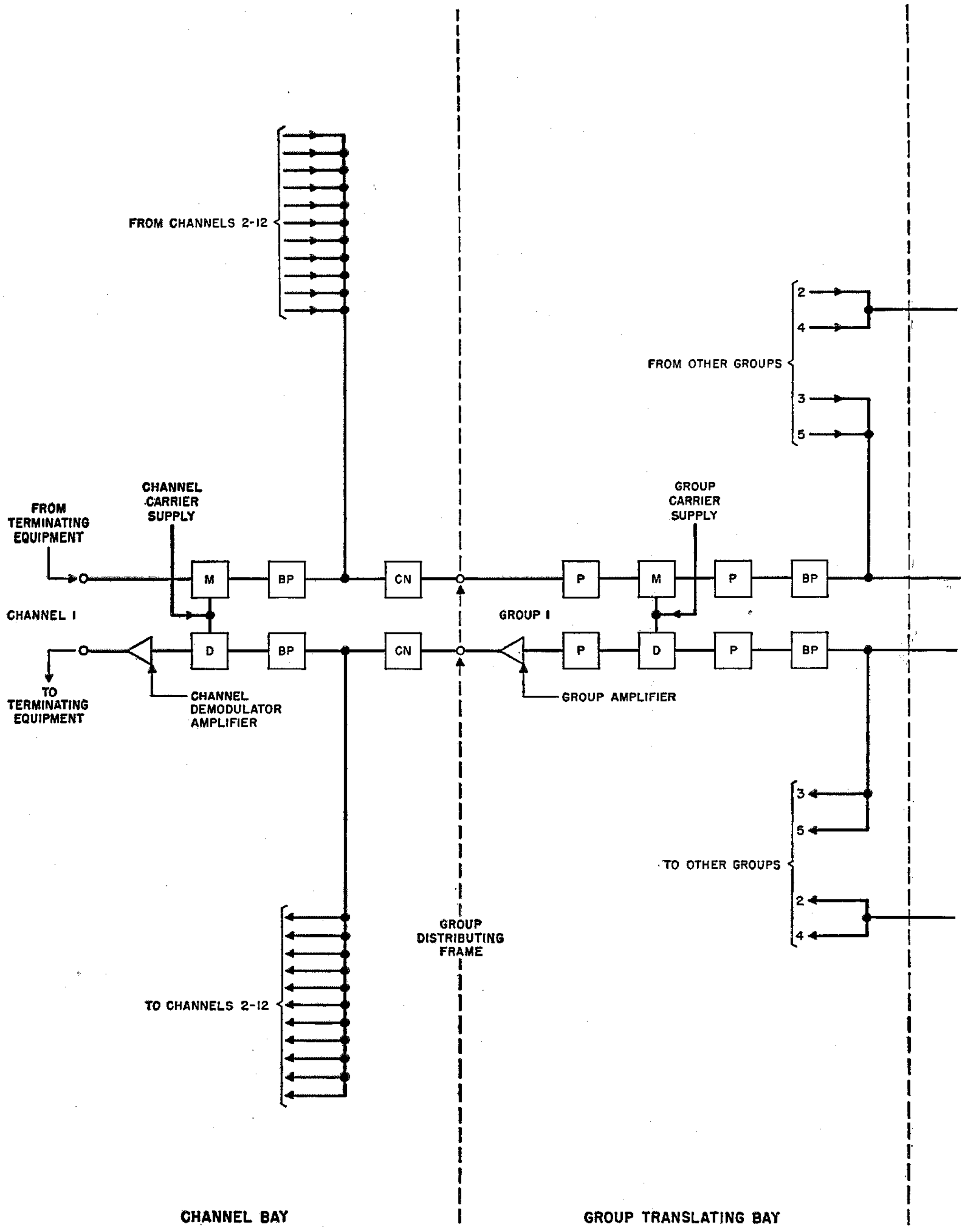
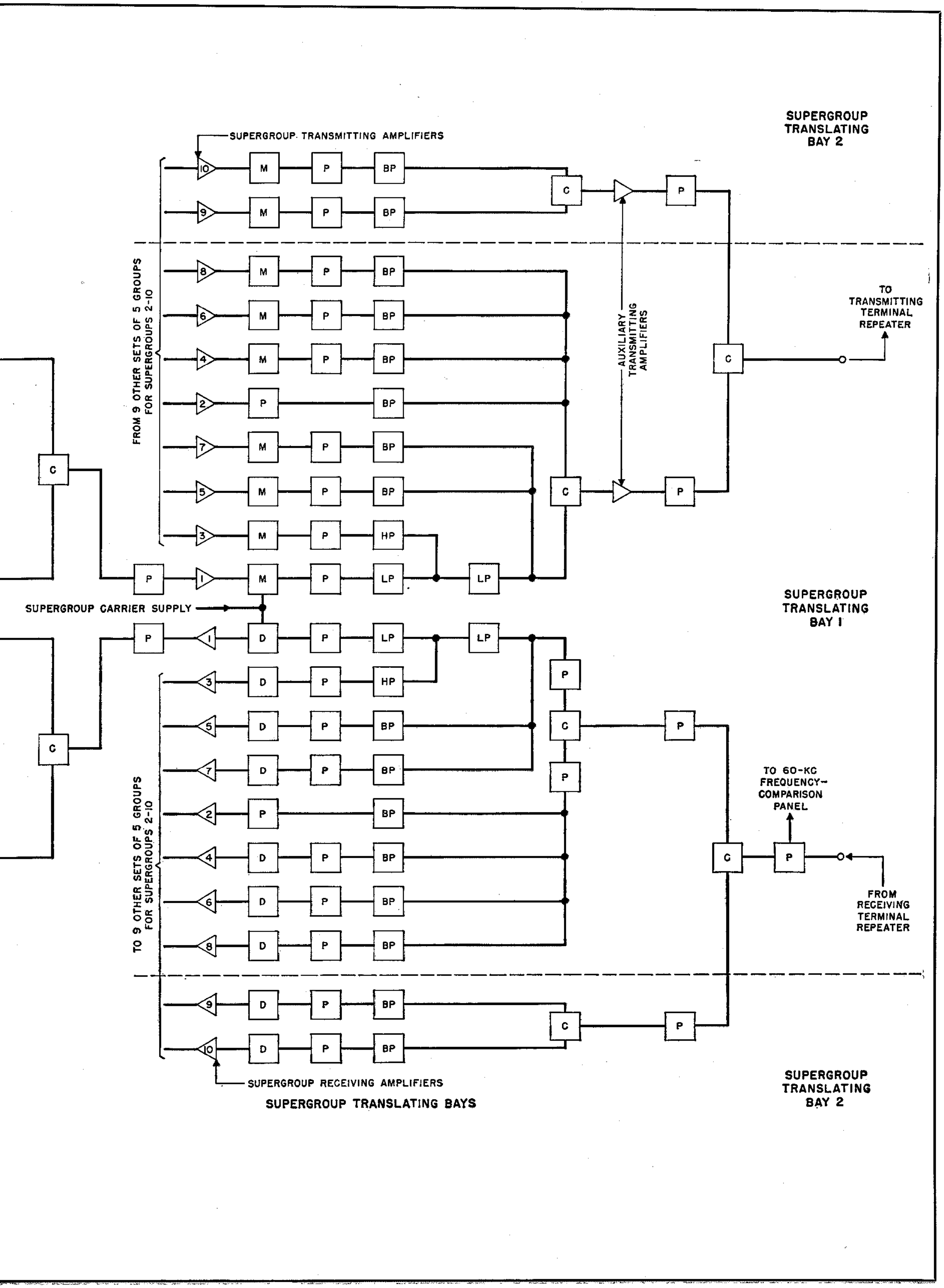


Fig. 4—Block diagram of terminal equipment, excluding terminal repeaters. The letters in the boxes indicate the following: C, Combining panel. BP, Band-pass filter. CN, Compensating network. D, Demodulator. HP, High-pass filter. LP, Low-pass filter. M, Modulator. P, Resistance pad.


SUPERGROUP TRANSLATING BAY 2
SUPERGROUP TRANSMITTING AMPLIFIERS
10
9
8
6
4
2
7
5
3
1
M
P
BP
HP
LP
C
FROM 9 OTHER SETS OF 5 GROUPS FOR SUPERGROUPS 2-10
AUXILIARY TRANSMITTING AMPLIFIERS
TO TRANSMITTING TERMINAL REPEATER
SUPERGROUP CARRIER SUPPLY
SUPERGROUP TRANSLATING BAY 1
D
TO 9 OTHER SETS OF 5 GROUPS FOR SUPERGROUPS 2-10
TO 60-KC FREQUENCY-COMPARISON PANEL
FROM RECEIVING TERMINAL REPEATER
SUPERGROUP RECEIVING AMPLIFIERS
SUPERGROUP TRANSLATING BAYS
SUPERGROUP TRANSLATING BAY 2

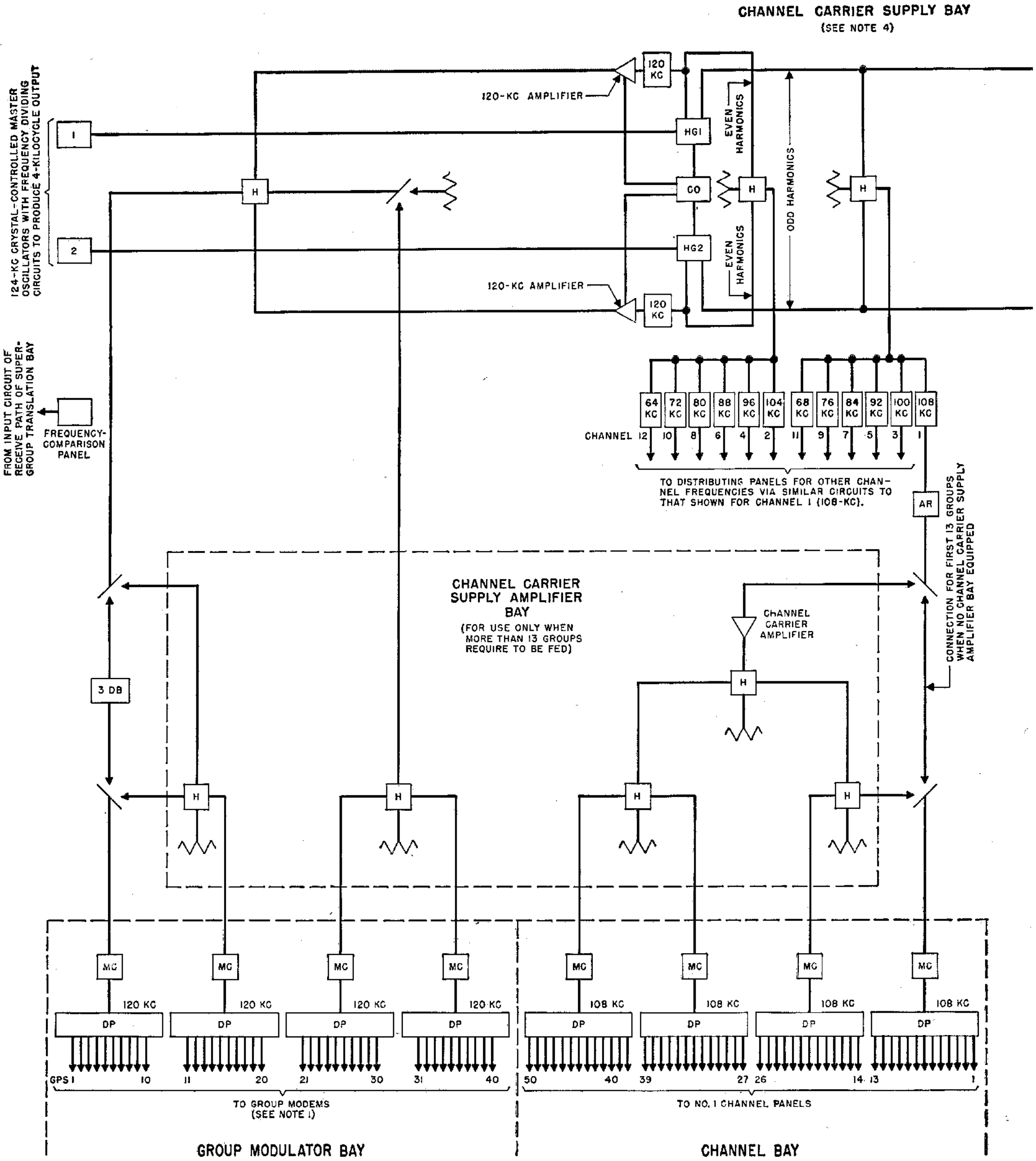
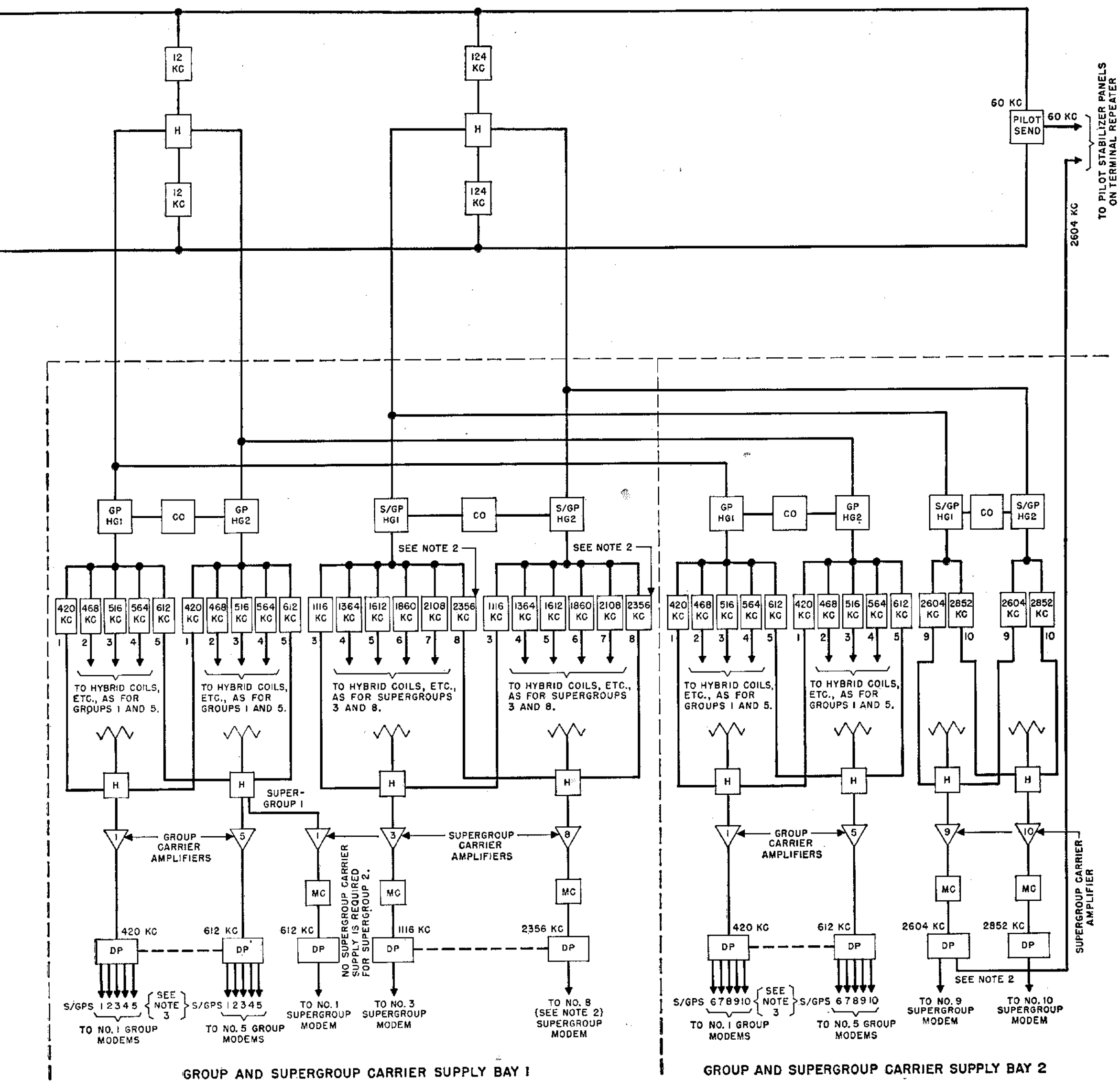

CHANNEL CARRIER SUPPLY BAY
(SEE NOTE 4)
124-KC CRYSTAL-CONTROLLED MASTER OSCILLATORS WITH FREQUENCY DIVIDING CIRCUITS TO PRODUCE 4-KILOCYCLE OUTPUT
1
2
120-KC AMPLIFIER
120 KC
HG1
HG2
CO
H
EVEN HARMONICS
ODD HARMONICS
FROM INPUT CIRCUIT OF RECEIVE PATH OF SUPER-GROUP TRANSLATION BAY
FREQUENCY-COMPARISON PANEL
64 KC
72 KC
80 KC
88 KC
96 KC
104 KC
68 KC
76 KC
84 KC
92 KC
100 KC
108 KC
CHANNEL 12 10 8 6 4 2 11 9 7 5 3 1
TO DISTRIBUTING PANELS FOR OTHER CHANNEL FREQUENCIES VIA SIMILAR CIRCUITS TO THAT SHOWN FOR CHANNEL 1 (108-KC).
AR
CONNECTION FOR FIRST 13 GROUPS WHEN NO CHANNEL CARRIER SUPPLY AMPLIFIER BAY EQUIPPED
CHANNEL CARRIER SUPPLY AMPLIFIER BAY
(FOR USE ONLY WHEN MORE THAN 13 GROUPS REQUIRE TO BE FED)
CHANNEL CARRIER AMPLIFIER
3 DB
MC
DP
120 KC
108 KC
GPS 1 10 11 20 21 30 31 40
50 40 39 27 26 14 13 1
TO GROUP MODEMS (SEE NOTE 1)
TO NO. 1 CHANNEL PANELS
GROUP MODULATOR BAY
CHANNEL BAY

Fig. 5—Block diagram of carrier supply. The letters in the boxes indicate the following: AR, Adjusting resistance. CO, Change-over (switching) panel. DB, Resistance pad of indicated attenuation in decibels. DP, Distributing panel. GP, Group. H, Hybrid coil. HG, Harmonic generator. KC, Band-pass filter of indicated midband frequency in kilocycles per second. MC, Matching coil. S/GP, Supergroup.

Note 1—A 120-kilocycle supply is only required for use with a group-modulator bay when the latter is used for extending a coaxial group direct to the first 12 channels of a 24-channel multi-conductor carrier system.

Note 2—When only 7 or fewer supergroups are required, the second group and supergroup carrier supply


12 KC
H
12 KC
124 KC
H
124 KC
60 KC
PILOT SEND
60 KC
2604 KC
TO PILOT STABILIZER PANELS ON TERMINAL REPEATER
GP HG1
CO
GP HG2
S/GP HG1
CO
S/GP HG2
SEE NOTE 2
SEE NOTE 2
GP HG1
CO
GP HG2
S/GP HG1
CO
S/GP HG2
420 KC
468 KC
516 KC
564 KC
612 KC
1116 KC
1364 KC
1612 KC
1860 KC
2108 KC
2356 KC
2604 KC
2852 KC
TO HYBRID COILS, ETC., AS FOR GROUPS 1 AND 5.
TO HYBRID COILS, ETC., AS FOR SUPERGROUPS 3 AND 8.
H
SUPER-GROUP 1
GROUP CARRIER AMPLIFIERS
SUPERGROUP CARRIER AMPLIFIERS
NO SUPERGROUP CARRIER SUPPLY IS REQUIRED FOR SUPERGROUP 2.
SUPERGROUP CARRIER AMPLIFIER
MC
DP
420 KC
612 KC
1116 KC
2356 KC
2604 KC
2852 KC
SEE NOTE 2
SEE NOTE 3
S/GPS 1 2 3 4 5
TO NO. 1 GROUP MODEMS
TO NO. 5 GROUP MODEMS
TO NO. 1 SUPERGROUP MODEM
TO NO. 3 SUPERGROUP MODEM
TO NO. 8 (SEE NOTE 2) SUPERGROUP MODEM
S/GPS 6 7 8 9 10
TO NO. 9 SUPERGROUP MODEM
TO NO. 10 SUPERGROUP MODEM
GROUP AND SUPERGROUP CARRIER SUPPLY BAY 1
GROUP AND SUPERGROUP CARRIER SUPPLY BAY 2

bay can be omitted. In this case, the 2356-kilocycle filter normally used for supergroup 8 is replaced by a 2604-kilocycle filter and the 2604-kilocycle pilot feed is taken from the amplifier normally associated with supergroup 8 instead of from that normally associated with supergroup 9.

Note 3—When 10 supergroups are initially equipped, each group-carrier amplifier on bay 1 would feed supergroups 1–5 and those on bay 2 would feed supergroups 6–10. The amplifiers, however, have a capacity for feeding a maximum of 8 supergroups, so that when bay 2 is not equipped, their load can be increased accordingly.

Note 4—The equipment not shown within the dashed line is mounted on the channel carrier supply bay.

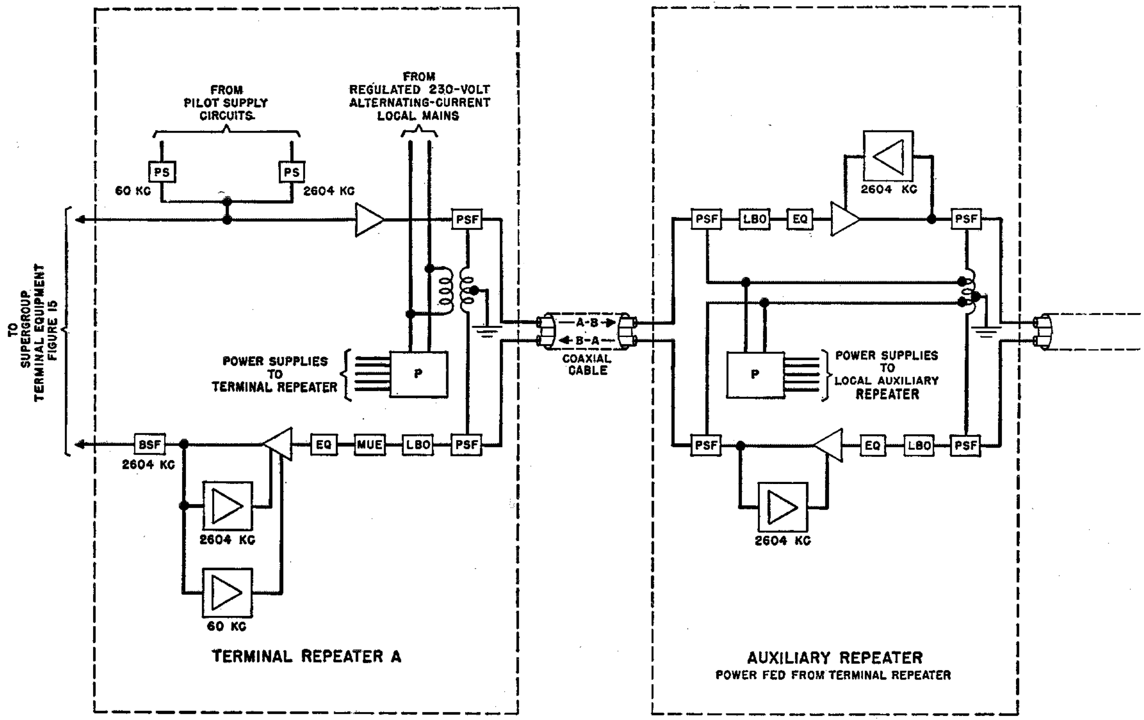
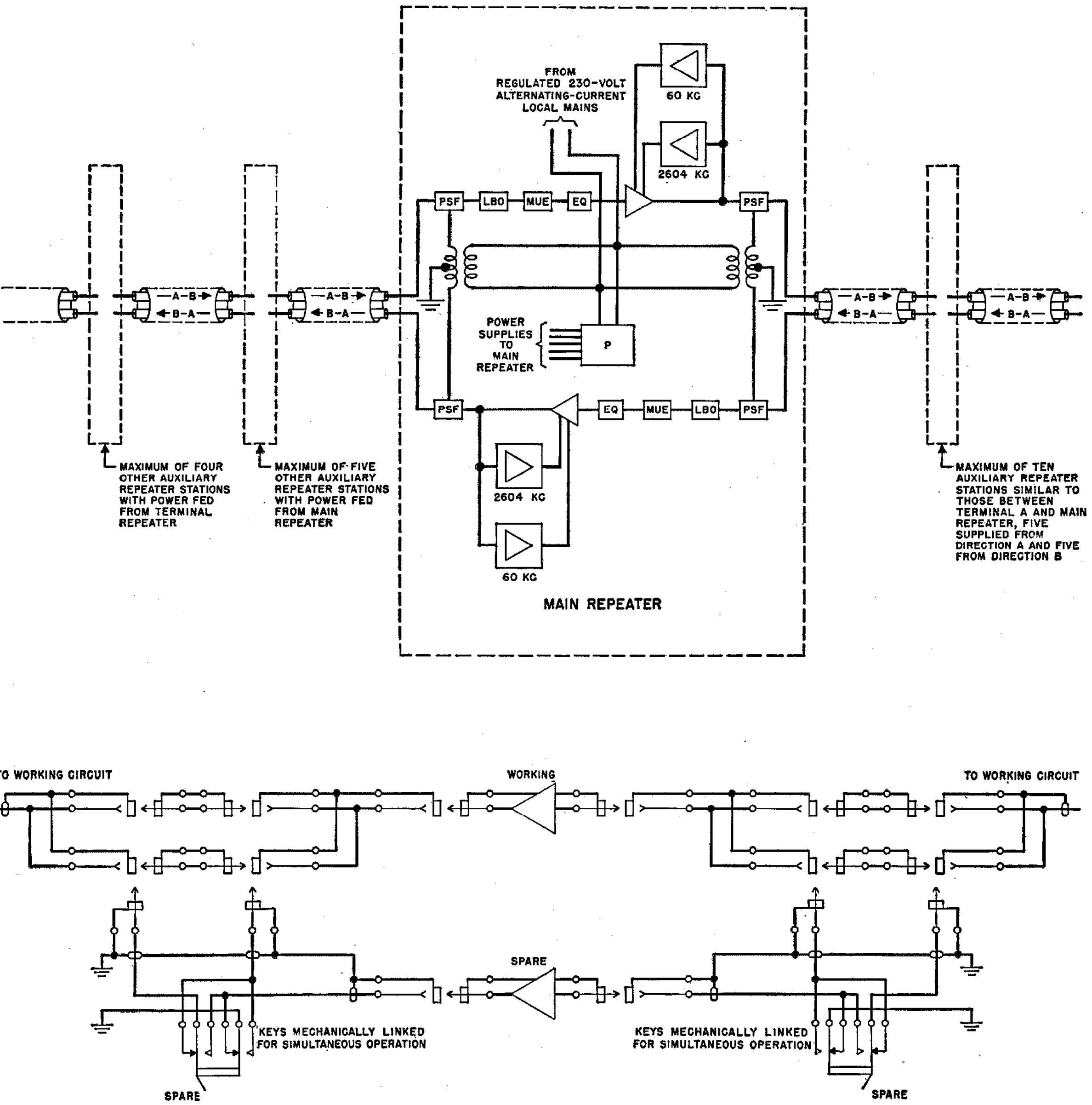

TO SUPERGROUP TERMINAL EQUIPMENT FIGURE 15
FROM PILOT SUPPLY CIRCUITS
FROM REGULATED 230-VOLT ALTERNATING-CURRENT LOCAL MAINS
PS
PS
60 KC
2604 KC
PSF
POWER SUPPLIES TO TERMINAL REPEATER
P
BSF
2604 KC
EQ
MUE
LBO
PSF
2604 KC
60 KC
TERMINAL REPEATER A
A-B
B-A
COAXIAL CABLE
2604 KC
PSF
LBO
EQ
PSF
P
POWER SUPPLIES TO LOCAL AUXILIARY REPEATER
PSF
EQ
LBO
PSF
2604 KC
AUXILIARY REPEATER
POWER FED FROM TERMINAL REPEATER

Fig. 6—Block diagram of line equipment including terminal main and auxiliary repeaters. The symbols in the boxes indicate the following: BSF, Band-stop filter. EQ, Line equalizer. LBO, line building-out network. MUE, Mop-up equalizer. P, Power unit. PS, Pilot stabilizer. PSF, Power-separating filter. ▷, Pilot regulator.

Note 1—Associated with each amplifier is a coaxial jack circuit, shown in sketch A, to enable a spare amplifier to be plugged in without interruption of service.


FROM REGULATED 230-VOLT ALTERNATING-CURRENT LOCAL MAINS
60 KC
2604 KC
PSF
LBO
MUE
EQ
PSF
A-B
B-A
POWER SUPPLIES TO MAIN REPEATER
P
PSF
EQ
MUE
LBO
PSF
2604 KC
60 KC
MAIN REPEATER
MAXIMUM OF FOUR OTHER AUXILIARY REPEATER STATIONS WITH POWER FED FROM TERMINAL REPEATER
MAXIMUM OF FIVE OTHER AUXILIARY REPEATER STATIONS WITH POWER FED FROM MAIN REPEATER
MAXIMUM OF TEN AUXILIARY REPEATER STATIONS SIMILAR TO THOSE BETWEEN TERMINAL A AND MAIN REPEATER, FIVE SUPPLIED FROM DIRECTION A AND FIVE FROM DIRECTION B
O WORKING CIRCUIT
WORKING
TO WORKING CIRCUIT
SPARE
KEYS MECHANICALLY LINKED FOR SIMULTANEOUS OPERATION
KEYS MECHANICALLY LINKED FOR SIMULTANEOUS OPERATION
SPARE
SPARE

SKETCH A
(SEE NOTE 1)

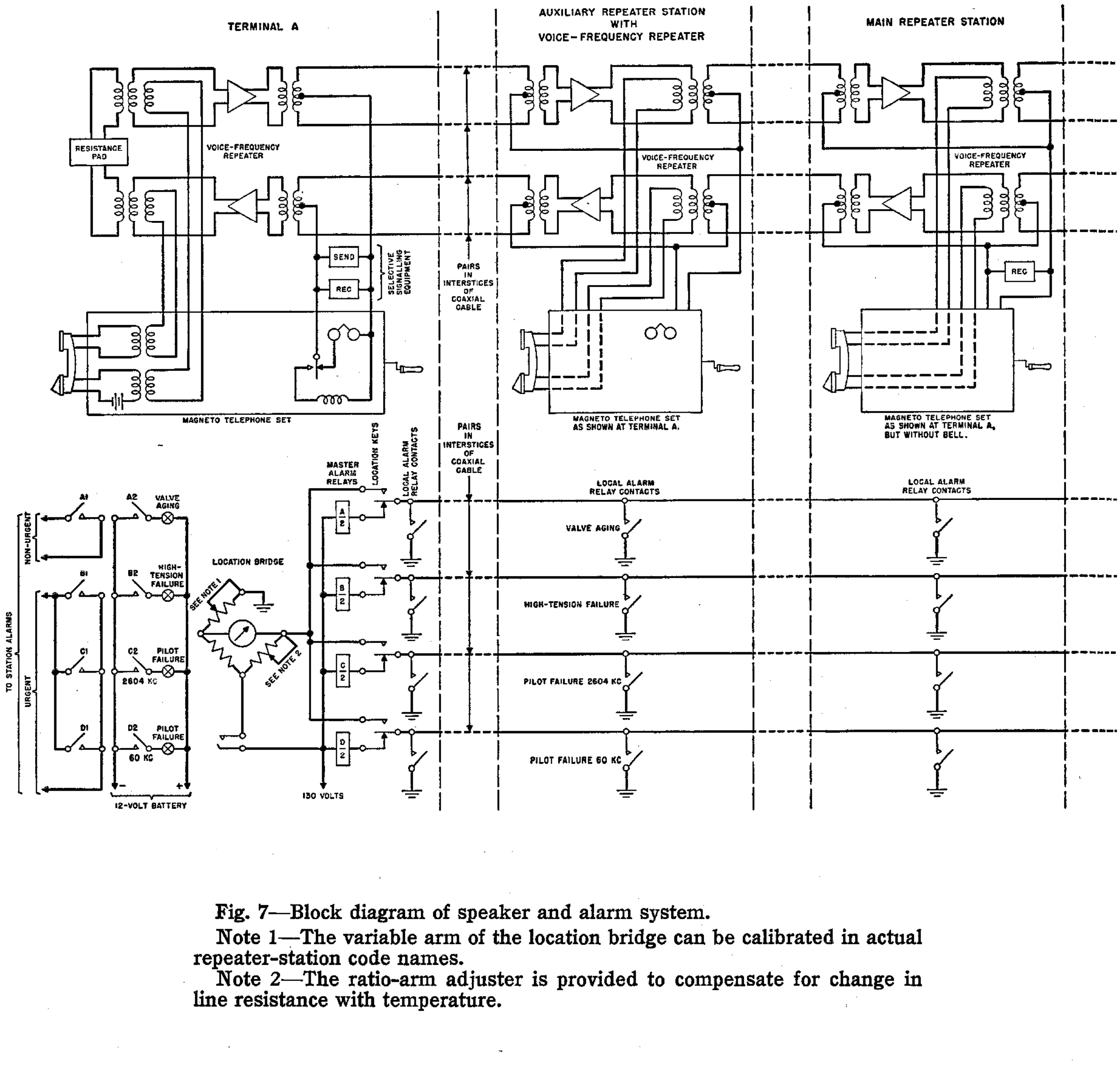
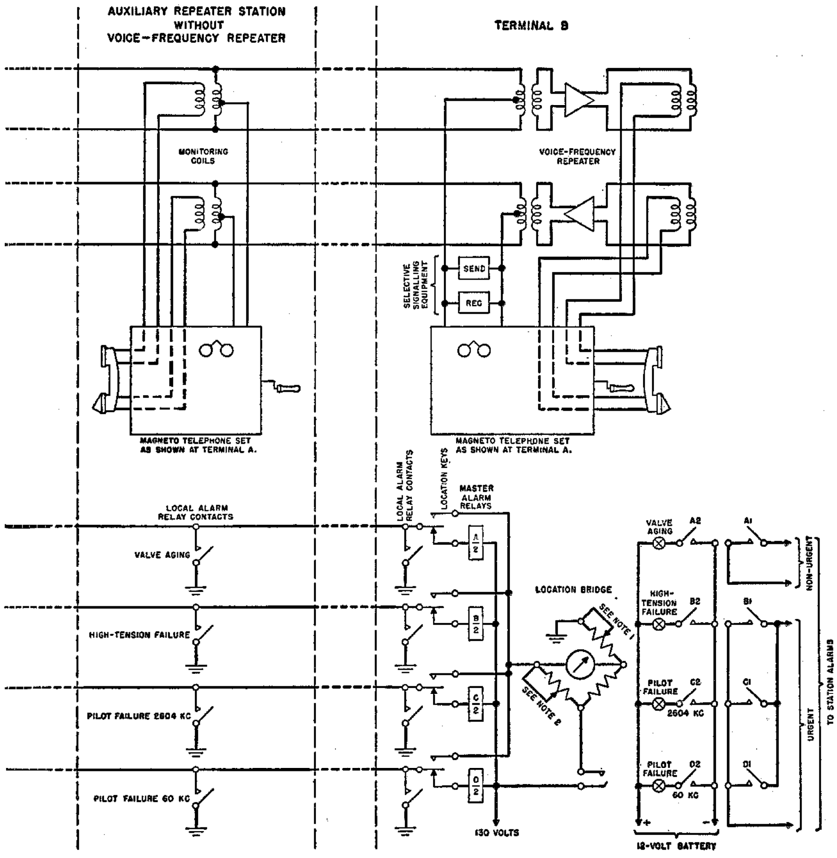


Fig. 7—Block diagram of speaker and alarm system.

Note 1—The variable arm of the location bridge can be calibrated in actual repeater-station code names.

Note 2—The ratio-arm adjuster is provided to compensate for change in line resistance with temperature.


AUXILIARY REPEATER STATION WITHOUT VOICE-FREQUENCY REPEATER
TERMINAL B
MONITORING COILS
VOICE-FREQUENCY REPEATER
SELECTIVE SIGNALLING EQUIPMENT
SEND
REC
MAGNETO TELEPHONE SET AS SHOWN AT TERMINAL A.
MAGNETO TELEPHONE SET AS SHOWN AT TERMINAL A.
LOCAL ALARM RELAY CONTACTS
VALVE AGING
HIGH-TENSION FAILURE
PILOT FAILURE 2604 KC
PILOT FAILURE 60 KC
LOCAL ALARM RELAY CONTACTS
LOCATION KEYS
MASTER ALARM RELAYS
A/2
B/2
C/2
D/2
130 VOLTS
LOCATION BRIDGE
SEE NOTE 1
SEE NOTE 2
VALVE AGING
A2
HIGH-TENSION FAILURE
B2
PILOT FAILURE
C2
2604 KC
PILOT FAILURE
D2
60 KC
+
−
12-VOLT BATTERY
A1
B1
C1
D1
NON-URGENT
URGENT
TO STATION ALARMS

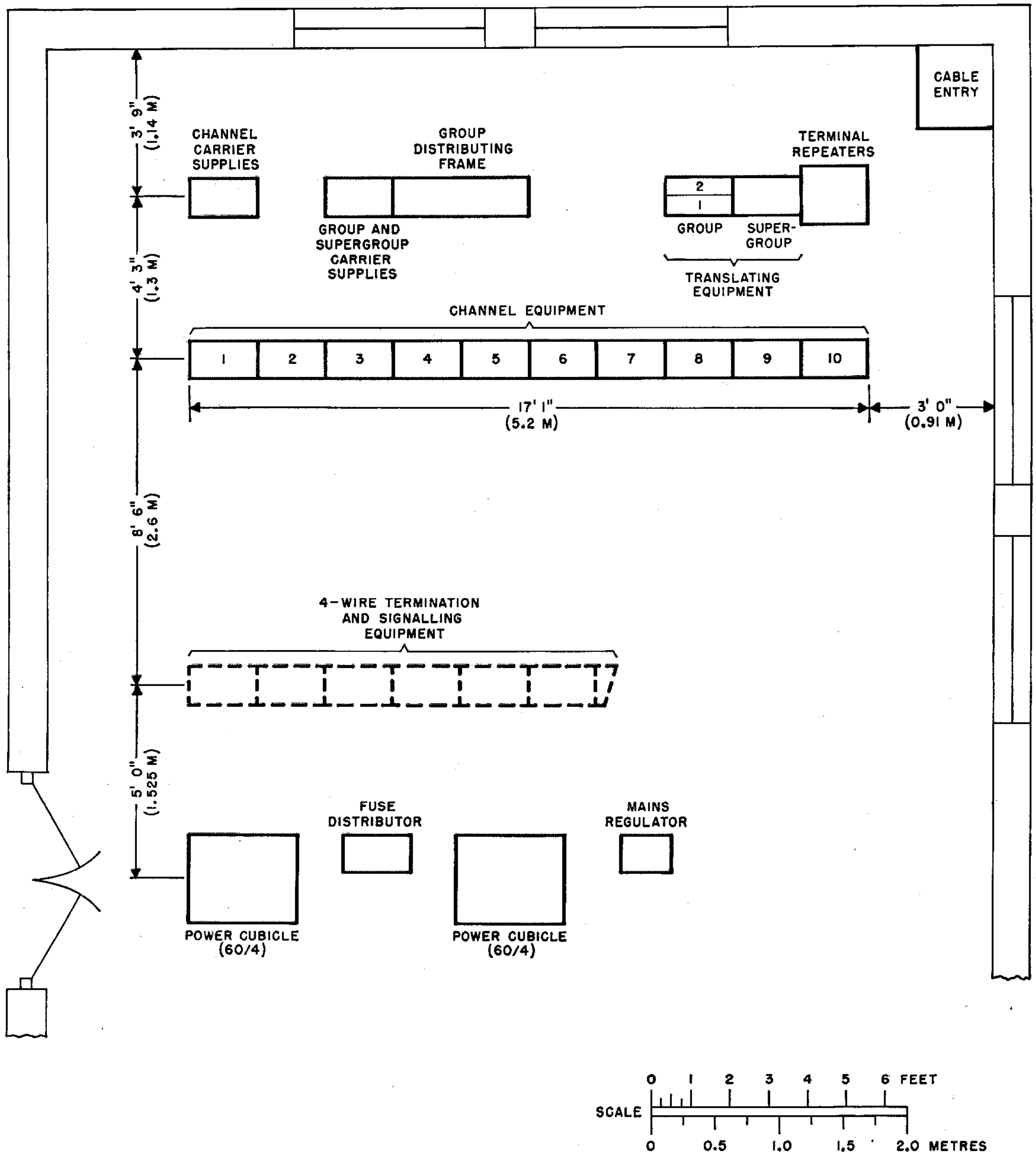
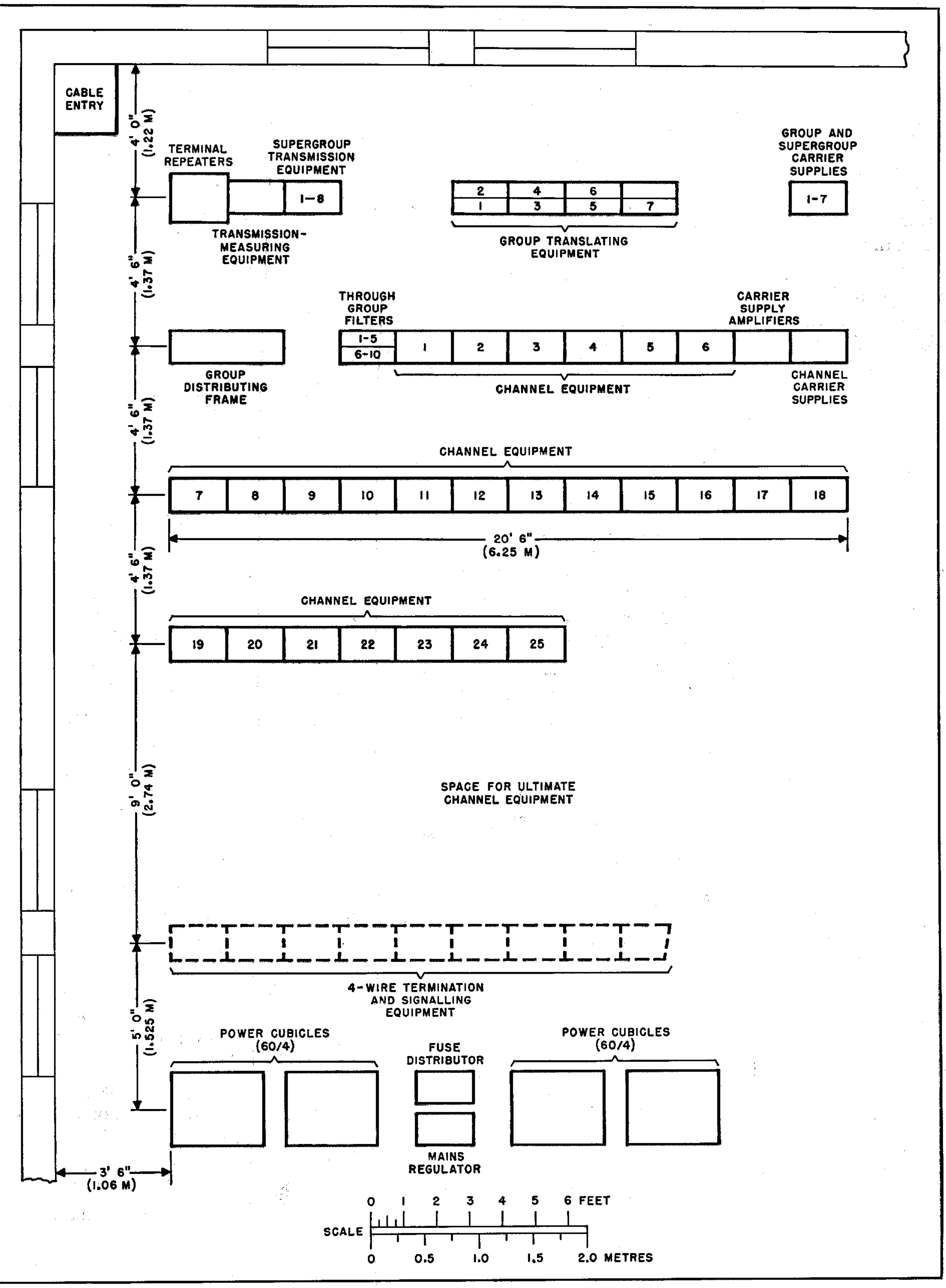


Fig. 8—Typical layouts for large (at right) and small (above) terminal stations, excluding standby power equipment. On each drawing, one of the power cubicles is intended to be reserved as a spare, the remainder having the capacity to supply all the equipment shown.

THESE LAYOUTS CATER FOR

	GROUPS TERMINATED AT STATION ILLUSTRATED		GROUPS FOR EXTENSION TO OTHER SYSTEMS	
	INITIALLY	ULTIMATELY	INITIALLY	ULTIMATELY
A	25	40	10	10
B	10	20	NIL	NIL


CABLE ENTRY
4' 0" (1.22 M)
TERMINAL REPEATERS
SUPERGROUP TRANSMISSION EQUIPMENT
1-8
TRANSMISSION-MEASURING EQUIPMENT
2 4 6
1 3 5 7
GROUP TRANSLATING EQUIPMENT
GROUP AND SUPERGROUP CARRIER SUPPLIES
1-7
4' 6" (1.37 M)
THROUGH GROUP FILTERS
1-5
6-10
1 2 3 4 5 6
CARRIER SUPPLY AMPLIFIERS
GROUP DISTRIBUTING FRAME
CHANNEL EQUIPMENT
CHANNEL CARRIER SUPPLIES
4' 6" (1.37 M)
CHANNEL EQUIPMENT
7 8 9 10 11 12 13 14 15 16 17 18
20' 6" (6.25 M)
4' 6" (1.37 M)
CHANNEL EQUIPMENT
19 20 21 22 23 24 25
9' 0" (2.74 M)
SPACE FOR ULTIMATE CHANNEL EQUIPMENT
4-WIRE TERMINATION AND SIGNALLING EQUIPMENT
5' 0" (1.525 M)
POWER CUBICLES (60/4)
FUSE DISTRIBUTOR
POWER CUBICLES (60/4)
MAINS REGULATOR
3' 6" (1.06 M)
0 1 2 3 4 5 6 FEET
SCALE
0 0.5 1.0 1.5 2.0 METRES

The carrier-supply circuit is shown in Fig. 5. The channel carrier-supply equipment for frequencies of 60 kilocycles for the pilot and 64–108 kilocycles for channel modulators is similar to that used on 12- and 24-channel systems. In addition, provision can be made to obtain 120 kilocycles for group-modulation purposes when required for transposing groups of a coaxial system directly to the 12–60-kilocycle range of a 12- or 24-channel system.

Since reliability of the whole terminal depends on the carrier supplies, vulnerable portions are duplicated, either completely, or by using additional valves; provision is made for automatic changeover from one set to the other in case of failure.

Each carrier-generating set consists of a crystal-controlled master oscillator with a fundamental frequency of 124 kilocycles. This is temperature compensated to give stability within ±1 part in 10 million, so that synchronisation is not necessary, although provision is made for occasional frequency comparison between terminals. The 124-kilocycle output is subdivided to produce 4 kilocycles to drive the harmonic generator supplying the harmonics of 4 kilocycles to the channel carrier-frequency-selecting filters and distributing points. The odd harmonics are provided by a saturated inductance, and the even harmonics by a rectifier circuit.

Odd harmonics of 12 kilocycles are produced in another frequency-generating panel to supply the 420-, 468-, 516-, 564-, and 612-kilocycle supplies for the group modulators and demodulators, the 612-kilocycle frequency also being used for number 1 supergroup. Odd harmonics of 124 kilocycles produced by a third generator provide the carrier frequencies for the remaining supergroup transposition processes. One of these frequencies—2604 kilocycles—is also used as a pilot to control the line amplifiers. Further details of the carrier-supply arrangements are given in the diagram.

2.5.2 *Coaxial-Line Equipment*

The line equipment is considered to include the line amplifying and equalising equipment, pilot-control apparatus, and power-separating filters at both terminals and repeaters, as well as the coaxial tubes themselves.

Fig. 6 is a schematic diagram showing various types of stations.

At the sending terminal, two pilot frequencies at 60 kilocycles and 2604 kilocycles are added to the outgoing supergroup signals and pass through the line amplifier combined with the power-separating filter to the coaxial tube. The level of each channel at this point is −1.15 neper (−10 decibels) relative to the transmitting toll-test-board level. At intermediate repeaters and at the distant receiving terminal, the incoming signals pass through the power-separating filter, building-out network, and equaliser, to the line amplifier; the gain and equalisation here are automatically controlled by the pilot currents through the use of thermistors. Thermistors are circuit elements in which the electrical resistance varies widely with change in temperature, and hence, with the amount of current passing through the thermistor.

Changes in line attenuation are mainly attributable to changes in temperature of the cable, and the "form" of the change is therefore predictable. This makes it possible to effect the necessary compensation by means of the 2604-kilocycle pilot only, at the majority of stations, while correcting for deviations at the low end of the frequency range by means of the 60-kilocycle pilot at certain repeaters and at terminals.

2.5.3 *Coaxial-System Power Supplies*

The channel, group, supergroup, and carrier-generating equipment at the terminals operate from 21- and 130-volt direct-current supplies. Single-phase alternating-current power at 50 cycles per second supplies the coaxial-line repeaters and, by feeding the power over the coaxial tubes themselves, it is possible to obviate the need for local power supplies at most of the intermediate repeater stations. Such stations are termed "auxiliary repeaters"; they are unattended, and in practice it is possible to have up to 10 such repeaters between power-feeding stations. Because so many circuits depend on coaxial systems, reliable standby power plant is needed, and it is customary to staff power-feeding stations which are, therefore, normally terminals and intermediate stations designated "main repeaters." It is usual to apply the 60-kilocycle-pilot correcting circuits as well as 2604-kilocycle control at these stations, and also to equip auxiliary equalisers. Also, since they are staffed, they are convenient stations at which to locate certain types of maintenance test gear.

The application of power to the coaxial cable at 650 volts involves the use of filters to separate the power- and telephone-transmission paths, and also special interlocking protection devices to prevent handling the dangerous parts of the circuit while the power is on.

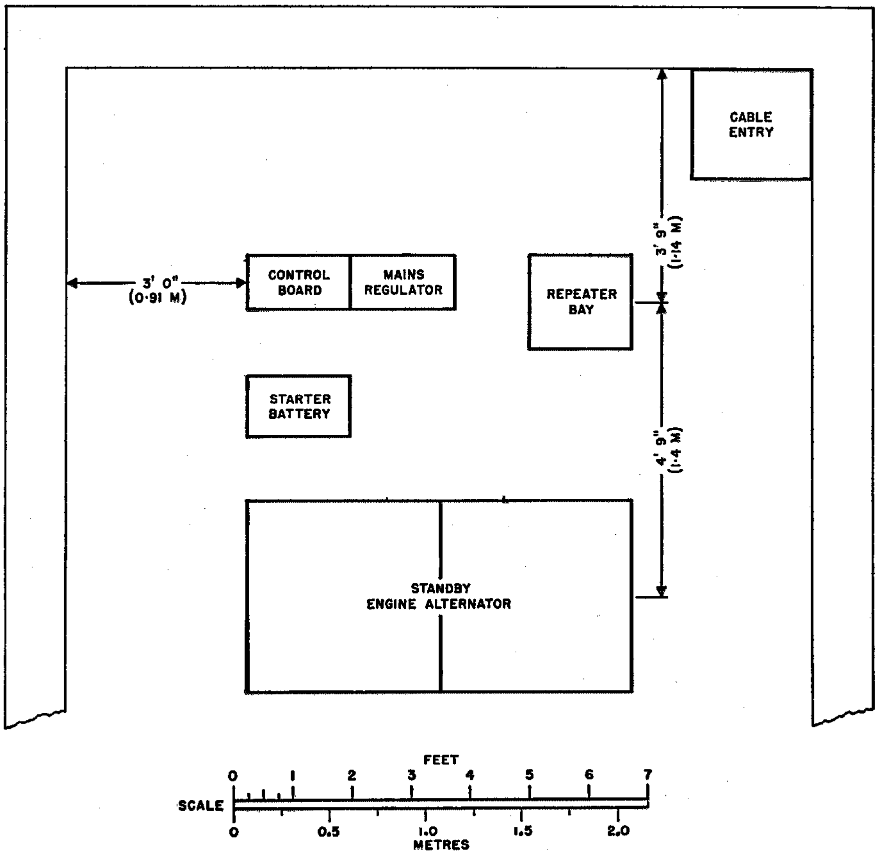


Fig. 9—Typical layout for main repeater station including standby power equipment.

2.5.4 Maintenance

To assist with inter-station maintenance, two pairs of the interstice quads normally laid up in the cable together with the coaxial tubes, are used to provide a supervisory speech circuit, and two other pairs are used for transmitting alarm conditions. The arrangement is shown in Fig. 7; faults indicated back to the control station include failure of power supply, deterioration of valves, and failure of 2604- and 60-kilocycle pilots. The bridge locating circuit enables the controlling station to identify the station from which the fault originates. In association with the above-mentioned supervisory facilities, specialised testing gear is also provided to maintain the system on a routine basis while it is in normal traffic operation or when faulty conditions occur.

2.5.5 Station Layout

Figs. 8A and 8B show typical station layouts for large and small coaxial terminals as indicated in the table. They include spare power cubicles for normal derivation of 21- and 130-volt direct-current supplies from the mains, but do not include spare standby plant for replacing the mains.

Fig. 9 shows a typical layout of a main repeater, and in this case the mains standby plant is included. The floor plan for an auxiliary repeater station is not shown, since this requires only one small bay of equipment, approximately 2-metres (6½-feet) high.

Various types of housing can be used for auxiliary repeater stations, including small metal-clad cupboards similar to lighting or tramway pillars which serve to enclose the bay.

The problem of housing equipment is nowadays of great importance because of the need for concentrating building materials and labour on the reconstruction of the large number of houses and other domestic buildings destroyed during the war. This problem has been aggravated by the lack of building during that period, and the result has been that considerably more attention has been paid to the reduction in size of equipment than would have been considered justified

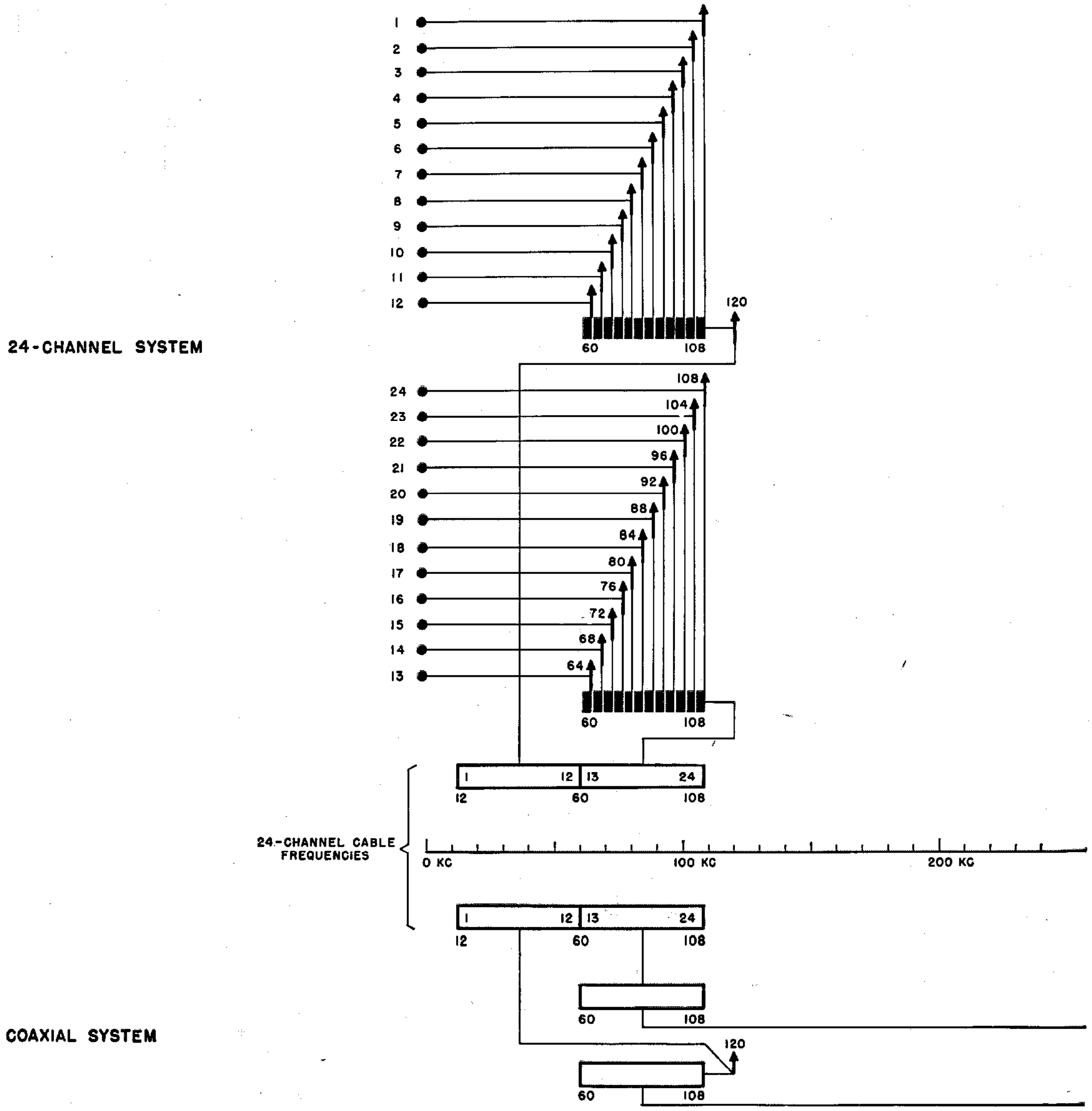
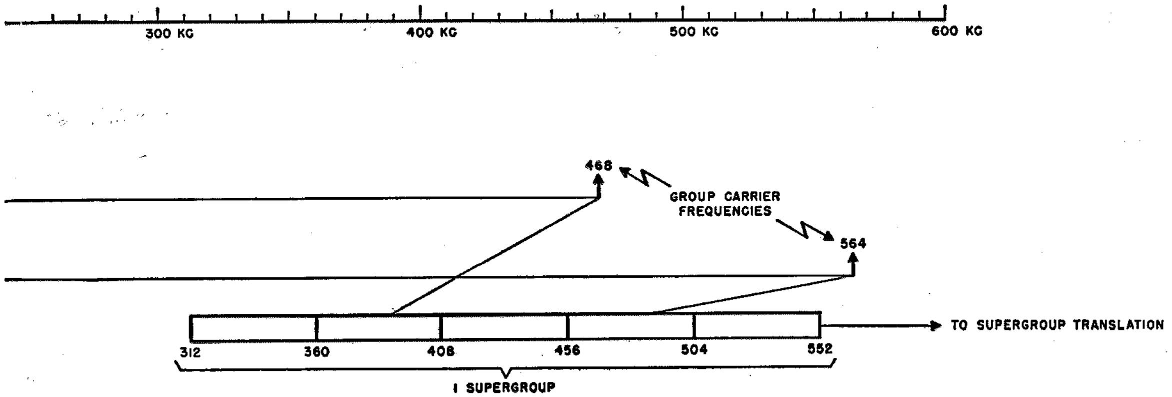


Fig. 10—Typical frequency allocation and routing of coaxial groups to 24-channel system.


300 KC
400 KC
500 KC
600 KC
468
GROUP CARRIER FREQUENCIES
564
TO SUPERGROUP TRANSLATION
312
360
408
456
504
552
1 SUPERGROUP

in the past. It is naturally important that reduction in size should not be accomplished at the expense of performance, and studies therefore start with this as a requirement. The channelling equipment offers more scope for reduction in size, and this reduction is well on the way to fulfillment. As is shown here, the high-frequency part of the repeater represents so small a proportion of the whole, that reduction in size must apply to the power plant or supervision facilities. These can only be simplified at the expense of modifying requirements, so that for the present little change is foreseen.

2.6 Interworking of Systems

So far, the main methods of obtaining first-grade telephone circuits have been discussed, but an important aspect from the point of view of a telephone administration working to build up a network is the degree to which standardisation of apparatus, circuits and operating and maintenance practices can be obtained. To achieve a closely welded overall system, it is obviously desirable that blocks of channels, quality provided by each channel, guaranteed carrier-supply systems, etc., all conduce to this end.

It has already been pointed out that the systems discussed here use the basic 12-channel group in the 60–108-kilocycle frequency range, and that this group is used to build up 12-channel, 24-channel, and coaxial systems; and it is also used for the 12-channel open-wire carrier telephone system. All the channel band-pass filters that control the quality give a similar performance and are made in large quantities, thereby reducing manufacturing costs. Fig. 10 shows how the frequency allocations of the 12-/24-channel systems, and the basic supergroup of the coaxial systems, permit flexibility and interworking. At junctions of multi-pair carrier systems, coaxial systems, or 12-channel open-wire systems, it is unnecessary to reduce channels to the voice-frequency range to interconnect on a group basis. Arrangements are made so that a coaxial terminal is normally broken down to the basic-group stage, and any such group is then available for routing at will (through a distribution frame) to the group stage of any of the following:

A. 12-channel carrier-on-cable system.
B. 24-channel carrier-on-cable system.
C. 12-channel open-wire system.
D. A second coaxial-cable system.
E. A voice-frequency terminal.

In the case of *A* to *D*, it is necessary to include a filter between the two connected systems, as shown in Fig. 11.

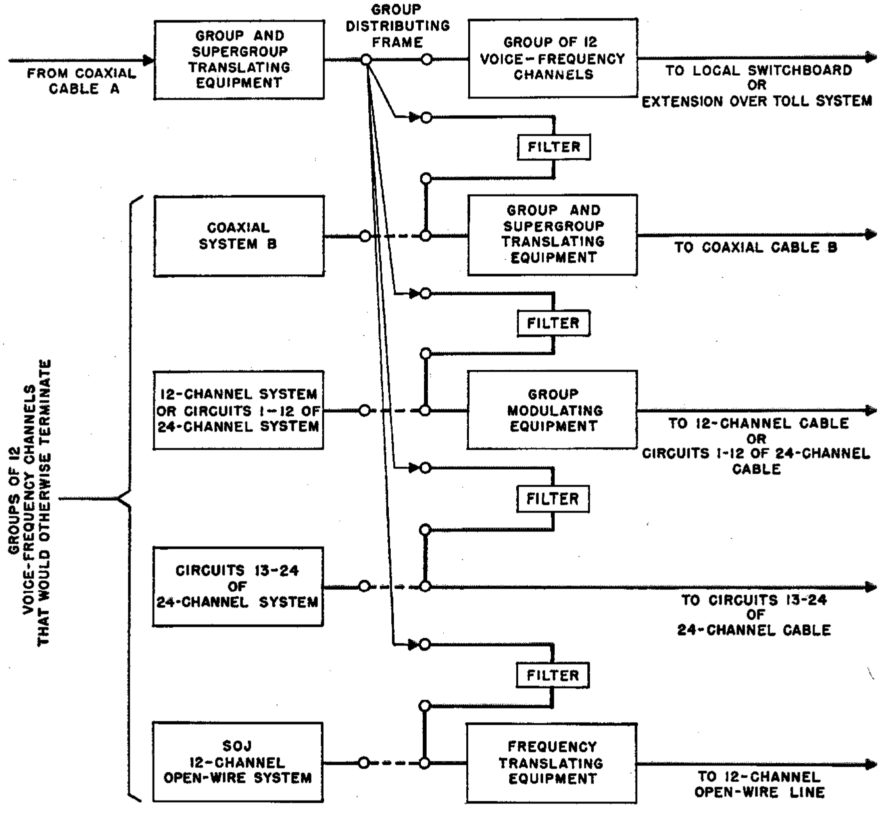

GROUP DISTRIBUTING FRAME
FROM COAXIAL CABLE A
GROUP AND SUPERGROUP TRANSLATING EQUIPMENT
GROUP OF 12 VOICE-FREQUENCY CHANNELS
TO LOCAL SWITCHBOARD OR EXTENSION OVER TOLL SYSTEM
FILTER
COAXIAL SYSTEM B
GROUP AND SUPERGROUP TRANSLATING EQUIPMENT
TO COAXIAL CABLE B
FILTER
12-CHANNEL SYSTEM OR CIRCUITS 1-12 OF 24-CHANNEL SYSTEM
GROUP MODULATING EQUIPMENT
TO 12-CHANNEL CABLE OR CIRCUITS 1-12 OF 24-CHANNEL CABLE
GROUPS OF 12 VOICE-FREQUENCY CHANNELS THAT WOULD OTHERWISE TERMINATE
FILTER
CIRCUITS 13-24 OF 24-CHANNEL SYSTEM
TO CIRCUITS 13-24 OF 24-CHANNEL CABLE
FILTER
SOJ 12-CHANNEL OPEN-WIRE SYSTEM
FREQUENCY TRANSLATING EQUIPMENT
TO 12-CHANNEL OPEN-WIRE LINE

Fig. 11—Diagram illustrating alternative routing of a single coaxial group.

2.7 Carrier on De-Loaded Cables

There is no doubt, in our view, that for multipair carrier working, the straightforward 2-cable system is best and most economical. Occasions can arise, however, where for the time being, at any rate, it is considered essential to de-load and use an existing cable or cables. The case of two such cables used on a go-and-return basis in the normal manner has been already treated by A. J. Jackman and R. A. Seymour[3]. In the special case where it is necessary to utilise one cable for transmission in both directions, the two directions are separated on a group-frequency basis, an untransposed basic group being transmitted in one direction and a transposed group in the other.

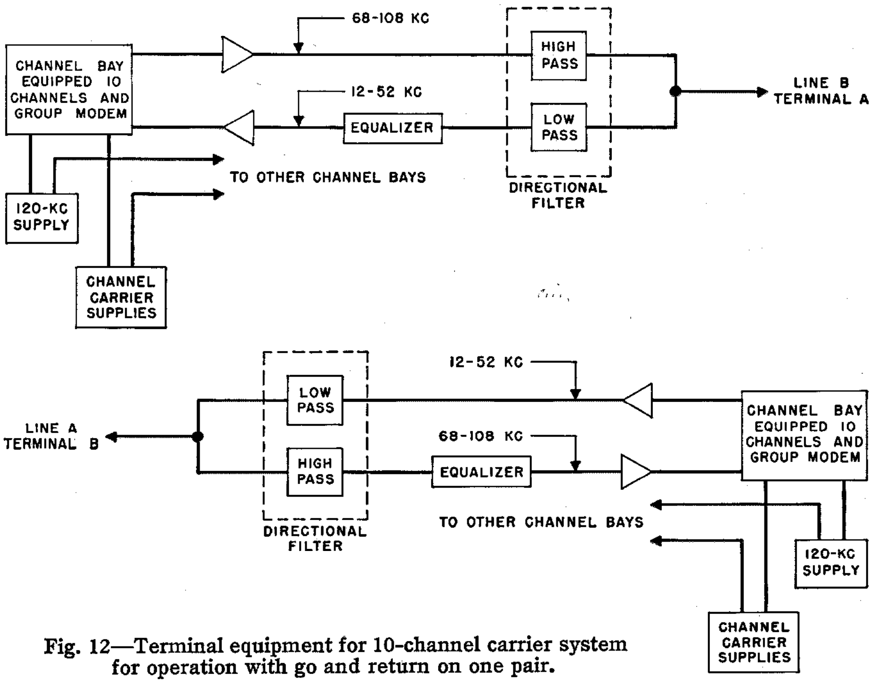


Fig. 12—Terminal equipment for 10-channel carrier system for operation with go and return on one pair.

2.8 12-Channel Carrier Using Separate Pairs

In this arrangement, 12 channels in the 12–60-kilocycle range are transmitted on one side circuit of a quad, and 12 in the 60–108-kilocycle range are transmitted in the other direction on the other side circuit. This method avoids rebalancing the cable, and is really a near approach to the standardised system, since repeaters are straightforward and no directional or separating filters are needed. In practice, 24-channel-system amplifiers and equalisers may be used, and at a later date when an additional cable can be laid, the same repeater equipment may be employed to double the capacity of the route.

2.9 10-Channel System on One Pair

By this method, any pair that can be made suitable from a crosstalk-balancing point of view may be made to carry 10 channels in each direction. An arrangement of the terminals is shown in Fig. 12. At one terminal, the unmodulated basic group is reduced to 10 channels by dropping the channels between 60 and 68 kilocycles, and is transmitted through a high-pass filter to line. At the distant terminal this is received through a high-pass filter, an equaliser, and an amplifier. No group modulation is required.

In the other direction, the 10 channels occupying the range of 12–52 kilocycles, which have been transposed in the normal way, are transmitted and received through the low-pass directional filter. The normal through repeater is shown in Fig. 13. This must utilise two directional filters to separate the two directions of transmission. The gap between 52 and 68 kilocycles is necessary to obtain adequate crossover attenuation of the filters, and also to obtain suitable impedance matching of the channels near the cut-off—an important point on long systems of this type. Fig. 13 also shows a frequency-transposing repeater, which may be used at a junction point of three cables to co-ordinate the frequencies in the different directions of transmission.

Comparing this scheme with the normal 4-wire 24-channel system, i.e., working with two cables, the 10-channel system does not appear at an advantage. Based on an equal number of channels, the 2-wire-system repeater is roughly about five times as expensive as the 24-channel-system repeater, because there are twice the number of amplifiers and equalisers (since only 10 channels are amplified in a group), and there is also the additional cost of the directional filters. Since there are more than twice the number of amplifiers for the same number of circuits, the power requirements are correspondingly increased. The equipment may also be of the order of three times the bulk, and hence the building-space requirement is correspondingly greater. The cost difference at terminals is relatively small, so that the overall economics of the two systems depend on the number of circuits required, the cost of the second cable, and the length of the circuits, i.e., the relative preponderance in cost of repeaters over terminals and cable. The general disadvantage of the 10-channel 2-wire system is that eventually, when a more economical 4-wire arrangement is realised, the directional filters are of no further use. For this reason the arrangement described in Section 2.8 is to be preferred.

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[3] A. J. Jackman and R. A. Seymour, "Application of 12 Circuit Carrier Telephony to Existing Cables," *The Post Office Electrical Engineers' Journal*, v. 34, p. 169; April, 1941.

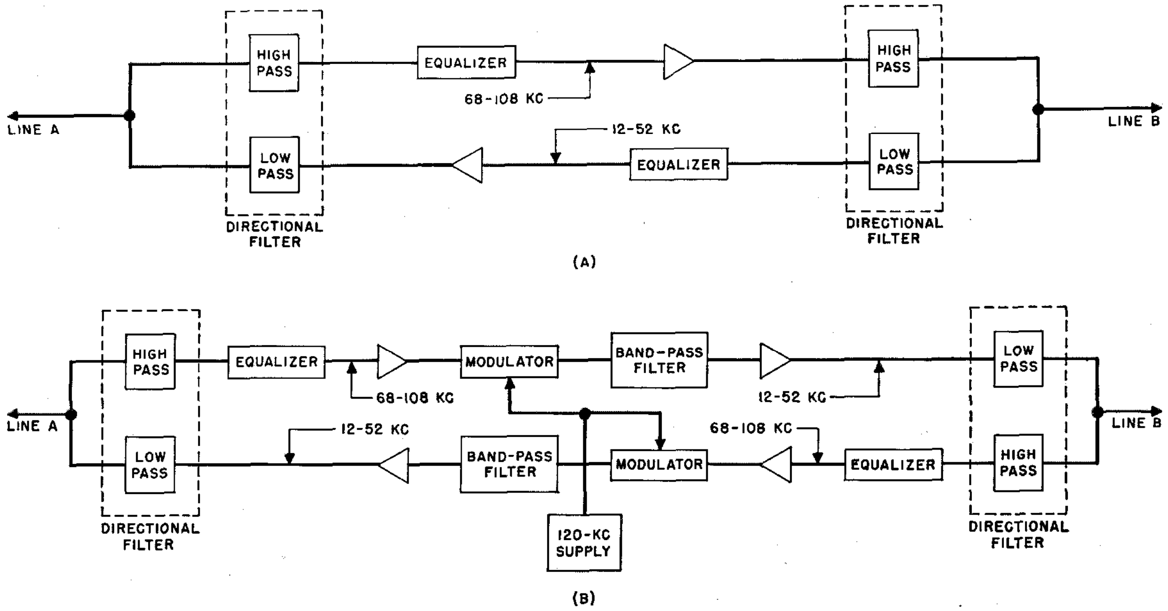

Fig. 13—(*A*)—Normal through repeater. (*B*)—Frequency-translating repeater for 10-channel carrier system for operation with go and return on one pair.

3. *Power and Supervision*

A point that becomes of major importance as equipment becomes smaller, and more and more channels can be carried by one amplifier, concerns what may be called auxiliary apparatus, notably that connected with power supply and supervision. In any multi-channel system, these tend to occupy considerably more space than the high-frequency parts of the equipment, and so present problems to which careful thought has had to be given.

Since in a coaxial-cable system there must exist a cable, it seems reasonable to expect that the cable itself can be used for these purposes, and this, in fact, turns out to be the case.

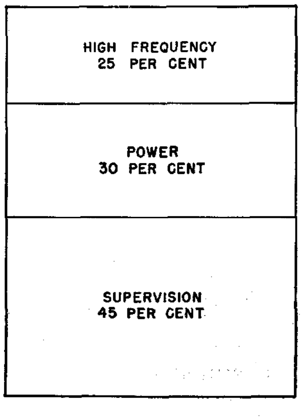

Fig. 14—Relative space occupancy of main divisions of coaxial repeater.

The power-supply problem is neatly solved by the transmission of power through the coaxial tubes themselves. The apparatus at the repeater station used in connection with power supply occupies approximately 30 per cent of the total volume of the repeater.

Supervision is a more difficult problem. When stations are unattended it is always necessary to provide means for indicating faults and for enabling visiting maintenance staff to converse between the station and the nearest control station. In the coaxial-cable system, by use of inexpensive pairs in the cable itself, supervision and maintenance circuits can easily be made available, but the volume occupied by these features in the coaxial repeater is approximately 45 per cent of the total.

Thus, as illustrated in Fig. 14, it will be seen that only 25 per cent of the volume of the repeater is occupied by the amplification and equalisation equipment. This point is stressed, as statements about the size of the repeater are frequently confused with statements about the size of the amplifier.

Only two types of fault indication are essential, and these can be designated urgent and non-urgent. Non-urgent faults can be dealt with at the time of the routine maintenance visit. (Presumably in the case of a cable there will always be a suitable road or track along which a vehicle can move, but for all ordinary purposes a bicycle is probably all that is necessary for use in unattended-repeater-station maintenance.) The urgent-alarm indication shows that a special visit is necessary to a particular repeater station. It is, of course, necessary to know which unattended station should be visited, and this can readily be found in the case of coaxial systems, as described earlier.

Urgent indication would be given only on complete failure of the repeater, and adequate spares that are automatically inserted obviously reduce the probability of the urgent-alarm feature being used. Of course, a non-urgent alarm would be given when a spare had come into use.

4. *Acknowledgment*

I wish to thank my many colleagues who have helped in the preparation of this paper, and in particular, Mr. A. M. Thornton.

Evaluation of Transmission Efficiency According to Hartley's Expression of Information Content*

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THE EFFICIENCY of any transmission system can be estimated as the ratio of the amount of information obtained in the reproduced message at the receiving end to the total amount of information that could have been transmitted theoretically in the propagation medium for the bandwidth and signal-to-noise ratio necessitated by the signals actually utilized. The amount of information is evaluated according to the expression given by R. V. L. Hartley for telegraphic signals, extended to the case of telephony and the presence of noise. The transmission efficiency thus defined is calculated for the main pulse-transmission systems (pulse-amplitude modulation, pulse-time modulation, pulse-count modulation, pulsed-frequency modulation) under certain assumptions. The various expressions obtained are helpful in the comparison between those systems, though it must not be overlooked that simplicity and cost of equipment will undoubtedly introduce other factors in the final choice for any given transmission problem.

• • •

1. General

To arrive at an evaluation of the amount of information conveyable through any type of transmission system, Hartley[1] started with the consideration of what would now be called pulse-amplitude-modulated (telegraphic) signals.

Let n_0 pulses be transmitted per second, each pulse having one of S discrete levels. The number of distinct sequences that can occur during a transmission time of t seconds is $S^{n_0 t}$. The measure of the amount of information that such a system is capable of transmitting is felt to be a function of this number of sequences. Barring any analysis of the psychological content of the message, which these signals are meant to convey, it becomes intuitive that this measure should be proportional to the time during which the transmission takes place. This determines at once the type of function to be used, so that, according to Hartley, the measure of the amount of information the signals can convey is expressed by

$$H = k_0 \cdot \log S^{n_0 t} = k_0 \cdot n_0 \cdot t \cdot \log S,$$

k_0 being a constant that, together with the base of the logarithms used, defines the value of a conventional unit of information.

This definition is easily extended to the case of the n_0 pulses belonging to n_c separate channels. The pulses pertaining to any one channel are divided into n_r subframes per second, and each subframe in its turn includes n pulses or digits. It is obvious that the total amount of information is then

$$H = k_0 \cdot n_c \cdot n_r \cdot n \cdot t \cdot \log S,$$

and is thus equal to n_c times the amount of information assigned to any individual channel.

These signals, however, will have to be transmitted through a certain path, which may include wire transmission lines, lumped circuits, and radio links. A certain number of errors may be introduced during that transmission. These errors may result from a number of causes, some originating in the physical properties of the path itself, such as signal distortion and interference as well as from thermal noise. Others may come from outside, such as interference from other transmissions or atmospherics. Moreover, the number of errors may depend on meteorological factors, such as temperature and humidity, leading to fluctuating distortion or fading phenomena. To secure a sufficiently correct reproduction of the "message" at the receiving end, it will thus be necessary to adopt a minimum value for the signal amplitude. If the limiting factor is thermal noise, it will establish a minimum signal-to-noise ratio at the receiving end for satisfactory communication. On the other hand, the type of signals and such properties of the transmission path as attenuation versus frequency and transmission delay or phase change versus frequency will determine the minimum frequency range or bandwidth for correct operation of the system.

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* Presented, New York Section, Institute of Radio Engineers, New York, New York, November 12, 1947.

[1] R. V. L. Hartley, "Transmission of Information," *Bell System Technical Journal*, v. 7, pp. 535–563; July, 1928.

The path over which the signals are transmitted will thus be characterized fundamentally by a minimum bandwidth and a minimum signal-to-noise ratio. An "ideal" type of "line" may be devised to serve as a basis for comparing different transmission systems. This ideal line will be endowed with the same bandwidth as the transmission path considered, but will present optimum properties as regards attenuation and phase response. That is to say, it will have a constant attenuation within its frequency range and infinite attenuation outside. The phase response will vary linearly with frequency within the bandwidth.

Such a line is, in effect, equivalent to an ideal low-pass filter with an abrupt cut-off frequency f_l. Any impulse of infinitely short duration but finite amplitude will produce a response proportional to

$$\frac{\sin 2\pi f_l \cdot t}{2\pi f_l \cdot t},$$

so that a series of such impulses separated by time intervals equal to $\frac{1}{2}f_l$ will generate independent instantaneous amplitudes which, theoretically at least, could be detected with negligible interference. However, the signal-to-noise ratio that must be used in the physical transmission path leads to a limitation on the number of separate levels that could otherwise be distinguished (those in the corresponding ideal line). Any signal amplitude S will be affected by noise; let $\bar{N}$ be the root-mean-square value of this noise, supposedly of thermal origin, i.e., having the normal random distribution. The probability of the deviation of the received amplitude being larger than $\bar{N}\sqrt{2}$ is small, about 5 percent. Thus, if we adopt $\bar{N}\sqrt{2}$ as the difference between two signal levels, we should have a small percentage of error due to noise. The total number of distinguishable levels on the ideal line is thus given by

$$\frac{S+\bar{N}\sqrt{2}}{\bar{N}\sqrt{2}}=1+\frac{S}{\bar{N}\sqrt{2}},$$

with a reasonable approximation. It follows that the amount of information transmittible on the ideal line is measured by

$$H_{lm}=k_0\cdot 2f_l\cdot t\cdot \log\left(1+\frac{S_l}{\bar{N}_l\sqrt{2}}\right).$$

It is suggested that the efficiency of the transmission system under examination be evaluated as the ratio of the amount of information recovered in the received message to this quantity H_{lm}, which is the maximum amount of information that can be transmitted on the corresponding ideal line.

In the case of telephony, however, the received message will generally consist of a continuous curve of amplitude versus time. This curve will be limited in frequency range; let f_m be the upper limit of this frequency band. It can be observed that, should this curve be sampled by means of equally spaced pulses at a rate of $2f_m$ per second, the amount of information contained in the series of pulses would be the same as for the continuous curve, since this latter curve could be extracted from the pulses by means of an ideal filter, and no other process would lead to any larger amount of information. Let S_m and $\bar{N}_m$ be the message amplitude and root-mean-square value of the message noise, respectively. The amount of information in the message is thus

$$H_m=k_0\cdot 2f_m\cdot t_m\cdot \log\left(1+\frac{S_m}{\bar{N}_m\sqrt{2}}\right).$$

According to the previous definition, the transmission efficiency of the system considered will be

$$\eta=\frac{H_m}{H_{lm}}=\frac{f_m t_m \log\left(1+\frac{S_m}{\bar{N}_m\sqrt{2}}\right)}{f_l t_l \log\left(1+\frac{S_l}{\bar{N}_l\sqrt{2}}\right)}.$$

It should be observed, however, that it is assumed here that the object of the transmission system is to reproduce with fidelity a continuous curve of amplitude versus time. No account is taken of the particular properties of the ear as the ultimate organ to be acted on, and all the following considerations suffer from that limitation.

The expression for η shows at once that all systems for which, in the ideal line, the product $f_l \cdot t_l$ remains constant and the signal-to-noise ratio unaltered, have the same efficiency. There can thus be an interchange between bandwidth and transmission time. A simple instance is the recording of a message to be transmitted, this record then being delivered to the line at a different speed from the recording speed.

Those systems for which $t_m = t_l$, as well as $f_m = f_l$, and $S_m/\bar{N}_m = S_l/\bar{N}_l$, have 100-percent efficiency. This is the case for single-sideband transmission over one message channel. When single-sideband transmission is used for a number of channels transposed in frequency (utilizing the so-called frequency-division method, as opposed to time division), the necessary line bandwidth is increased by the amount required for the separation of the different message channels by band-pass filters. This lowers the efficiency to about 85 percent when quartz filters of modern design are utilized.

The most prominent transmission systems are analyzed below in the light of the foregoing considerations. Frequency modulation (FM), pulse-time modulation (PTM), pulse-count modulation (PCM), pulse-amplitude modulation (PAM), and pulsed-frequency modulation (PFM) are included. It is assumed in all cases that a high signal-to-noise ratio is required for the message, and a value of 60 decibels has been chosen as being conventional.

2. *Frequency-Modulation Systems*

Frequency modulation is too well known to require any description. Let n_c channels be transmitted simultaneously, the frequency range of each channel being f_m. The information in the reproduced message will be

$$H_m = 2k_0 \cdot n_c \cdot f_m \cdot t \cdot \log\left(1 + \frac{S_m}{\bar{N}_m\sqrt{2}}\right).$$

The ideal-line characteristics are easily determined. The modulation frequency range will be somewhat greater than $n_c \cdot f_m$, say, $1.2n_c \cdot f_m$ as explained above, to allow for separation of the channels by means of filters. Let m be the modulation index. The minimum line-bandwidth for acceptable distortion is generally considered to be

$$2(m+1) \times 1.2n_c \cdot f_m.$$

The information content transmittible over the line will be

$$H_{lm} = 4k_0(m+1) \times 1.2n_c \cdot f_m \cdot t \cdot \log\left(1 + \frac{S_l}{\bar{N}_l\sqrt{2}}\right),$$

resulting in the following expression for the transmission efficiency:

$$\eta = \frac{1}{2.4(m+1)} \frac{\log\left(1 + \frac{S_m}{\bar{N}_m\sqrt{2}}\right)}{\log\left(1 + \frac{S_l}{\bar{N}_l\sqrt{2}}\right)}.$$

Now the relation between signal-to-noise ratios for the message and the line is known to be

$$\frac{S_l}{\bar{N}_l\sqrt{2}} = \frac{1}{m\sqrt{3}} \frac{S_m}{\bar{N}_m\sqrt{2}}.$$

This relation holds as long as $S_l/\bar{N}_l\sqrt{2}$ is sufficiently greater than 1 (say, 3), so that the signal remains above the so-called frequency-modulation threshold. As we assume $20 \log (S_m/\bar{N}_m\sqrt{2})$ to be equal to 60; this means that the validity of the expression for η is limited to values of m smaller than $1000/3\sqrt{3}$, or 192.

Up to a value of $m = 100$, the units under the logarithms can be neglected without appreciably altering the results, so that

$$\eta = \frac{1.25}{(m+1)(2.76 - \log m)}.$$

The results are shown in Table 1.

TABLE 1

TRANSMISSION EFFICIENCY FOR FREQUENCY-MODULATION SYSTEMS

m	η	m	η
1	0.23	7	0.080
2	0.17	8	0.075
3	0.14	9	0.070
4	0.12	10	0.065
5	0.10	50	0.020
6	0.09	100	0.016

3. *Pulse-Time-Modulation Systems*

For pulse-time modulation, the signals constitute a series of pulses which, in the absence of modulation, are of equal height, equal width, and equally spaced in time. This series is divided into n_c interlaced subseries, n_c being the number of channels. With the introduction of modulation, each pulse is displaced in time proportionally to the instantaneous amplitude of the modulating message. Means are provided at the receiving end to separate the subseries and convert the time modulation back into the original continuous curve of amplitude versus time that constituted the message.[2]

The message information content per channel is, therefore,

$$H_m = 2k_0 \cdot f_m \cdot t \cdot \log\left(1 + \frac{S_m}{\bar{N}_m\sqrt{2}}\right).$$

The ideal-line characteristics are determined as follows.

The frequency range will be such as to keep the interference between any one pulse and those following less than the difference between two distinguishable levels on the line.

Let τ be the pulse duration and $\pm d$ be the maximum time displacement; the number of channels is

$$n_c = \frac{1}{2f_m\tau\left(1 + \frac{2d}{\tau}\right)}.$$

To simplify, no account is taken of any marking pulse inserted for the separation of the channels.

The relation between the signal-to-noise ratios in the message and the line is found to be[2]

$$\frac{S_m}{\bar{N}_m\sqrt{2}} = \frac{d}{g}\,\frac{S_l}{\bar{N}_l\sqrt{2}},$$

g being the rise time of the signal, which is supposed to be approximately trapezoidal. A reasonable assumption is $g = \tau/2$. Then,

$$\frac{S_l}{\bar{N}_l\sqrt{2}} = \frac{\tau}{2d}\,\frac{S_m}{\bar{N}_m\sqrt{2}}.$$

In the following, d/τ will be kept under a value that will bring $S_l/\bar{N}_l\sqrt{2}$ below 3 (i.e., $d/\tau < 166$). Let this quantity be called m, and a value of 60 decibels be assumed for the message signal-to-noise ratio. The units under the logarithms in the expression for η can be neglected. Let the bandwidth of the line be called B; then finally,

$$\eta = \frac{1}{2B\tau(2m+1)}\,\frac{\log \frac{S_m}{\bar{N}_m\sqrt{2}}}{\log \frac{1}{2m}\frac{S_m}{\bar{N}_m\sqrt{2}}}.$$

The quantity $B\tau$ can now be determined with the condition of limited interference between successive pulses, assuming an exponential law for the attenuation of the signal with time on the line and a time constant of $1/\pi B$:

$$e^{\pi B\tau} > \frac{1}{2m}\,\frac{S_m}{\bar{N}_m\sqrt{2}},$$

$$B\tau = \frac{2.3}{\pi}\log\frac{1}{2m}\,\frac{S_m}{\bar{N}_m\sqrt{2}}.$$

For a message signal-to-noise ratio equal to 60 decibels, the transmission efficiency is expressed by

$$\eta = \frac{2}{(2m+1)(2.7 - \log m)^2}.$$

The results are shown in Table 2, which gives values for η when $g = \tau/2$ and $\tau/3$.

TABLE 2

TRANSMISSION EFFICIENCY FOR PULSE-TIME-MODULATION SYSTEMS

m	η ($g=\tau/2$)	η ($g=\tau/3$)	m	η ($g=\tau/2$)	η ($g=\tau/3$)
0.5	0.11	0.13	6	0.042	—
1	0.09	0.11	7	0.039	—
2	0.07	—	8	0.036	—
3	0.057	—	9	0.034	—
4	0.050	0.06	10	0.023	0.04
5	0.045	—			

4. *Pulse-Count-Modulation Systems*

Pulse-count modulation, as first described by A. H. Reeves in United States and French patents,[3] makes use of three successive operations: sampling, quantizing of the sampled amplitudes, and transmission of the quantized amplitudes by means of a code similar to that utilized at the beginning of this paper to introduce Hartley's method of measuring the amount of information conveyable in transmission systems. Quantizing means that the amplitude range of the modulating signal is divided into a finite number of discrete levels. Any sampled amplitude will fall between two successive levels; only that level closer to the actual signal amplitude is transmitted.[4-6]

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[2] E. M. Deloraine and E. Labin, "Pulse Time Modulation," *Electrical Communication*, v. 22, n. 2, pp. 91–98; 1944.

[3] A. H. Reeves, United States patent 2,272,070; February 3, 1942. French patent 852,183; October 23, 1939.

Let m be the number of levels used; the message information content, where n_c channels are transmitted, is

$$H_m = 2k_0 \cdot n_c \cdot f_m \cdot t \cdot \log m,$$

f_m being the frequency range of each individual channel.

Let the code comprise ν digits, each having s discrete values. This means that the number of levels is $m = s\nu$. The minimum bandwidth for the line can be shown[4] to be $\nu n_c f_m$, so that the transmission efficiency is

$$\eta = \frac{1}{\nu} \frac{\log m}{\log\left(1 + \frac{S_l}{\bar{N}_l\sqrt{2}}\right)}.$$

As $\nu = \log m / \log s$, the transmission efficiency can also be written

$$\eta = \frac{\log s}{\log\left(1 + \frac{S_l}{\bar{N}_l\sqrt{2}}\right)}.$$

When the code used is binary, $s = 2$, the relation between signal-to-noise ratio in the line and message can be shown to be

$$20 \log \frac{S_m}{\bar{N}_m\sqrt{2}} = 2.2 \left(\frac{S_l}{\bar{N}_l\sqrt{2}}\right)^2,$$

provided the number of digits in the code is sufficiently large, exceeding, say, 3 or 4.

Assuming that, as for the systems previously considered, a signal-to-noise ratio of 60 decibels is required for the message, the transmission efficiency of pulse-count-modulation systems is found to be approximately 0.4, and is independent of the number of digits in the code provided it is larger than 3 or 4.

This means that whatever the amount of distortion specified for the reproduced message (which determines the number of levels m) may be, the pulse-count-modulation transmission efficiency, as defined in the text, remains the same for a binary system of coding.

5. Pulse-Amplitude-Modulation Systems

With pulse-amplitude modulation transmitted over a line where transient response varies exponentially with time, and assuming n_c channels to be transmitted, each f_m cycles wide, the line bandwidth B will be conventionally determined by limiting to less than the minimum distinguishable level $\bar{N}_l\sqrt{2}$ the effect of interference produced by a pulse of maximum amplitude on a succeeding pulse.

This gives

$$e^{\pi B\tau} = \frac{S_l}{\bar{N}_l\sqrt{2}},$$

τ being the pulse duration, equal to $\frac{1}{4} n_c f_m$.

The relation between signal-to-noise ratios in message and line is, for full modulation,

$$\frac{S_l}{\bar{N}_l\sqrt{2}} = 2\left(\frac{f_m}{B}\right)^{\frac{1}{2}} \frac{S_m}{\bar{N}_m\sqrt{2}},$$

so that B/f_m is determined by

$$\frac{1}{N_c} \cdot \frac{B}{f_m} = \frac{3.2}{\pi} \log\left[2\left(\frac{f_m}{B}\right)^{\frac{1}{2}} \frac{S_m}{\bar{N}_m\sqrt{2}}\right].$$

Once B/f_m is found, the transmission efficiency can be calculated according to definition:

$$\eta = \frac{3.2}{\pi}\left(\frac{n_c f_m}{B}\right)^2 \log \frac{S_m}{\bar{N}_m\sqrt{2}}.$$

Let the message signal-to-noise ratio be 60 decibels as before; the two equations giving B/f_m and η are

$$\frac{1}{n_c} \frac{B}{f_m} = \frac{3.2}{\pi}\left(3.3 - 0.5 \log \frac{B}{f_m}\right),$$

$$\eta = \frac{27.6}{\pi}\left(\frac{n_c f_m}{B}\right)^2.$$

The results are listed in Table 3. The number of channels has been limited to a value for

TABLE 3

TRANSMISSION EFFICIENCY FOR PULSE-AMPLITUDE-MODULATION SYSTEMS

n_c	B/f_m	$B/n_c f_m$	η
1	8.3	8.3	0.13
20	130	6.5	0.21
50	300	6.0	0.24
100	555	5.5	0.28
400	1950	4.9	0.37

---

[4] D. D. Grieg, "Pulse-Count Modulation," *Electrical Communication*, v. 24, pp. 287–296; September, 1947.

[5] H. S. Black and J. O. Edson, "PCM Equipment," *Electrical Engineering*, v. 66, pp. 1123–1125; November, 1947.

[6] W. M. Goodall, "Telephony by Pulse Code Modulation," *Bell System Technical Journal*, v. 26, pp. 395–409; July, 1947.

which the signal-to-noise ratio in the line remains large enough to allow for substantially 100-percent modulation.

6. Pulsed-Frequency-Modulation Systems

By pulsed-frequency modulation is meant a system utilizing a series of pulses of equal height, equal width, with fixed position in time, and the information being transmitted by frequency modulating the high-frequency carrier wave.

The required line bandwidth is then approximately

$$B_0+2(m+1)f_m,$$

B_0 being the additional bandwidth needed for the correct transmission of the pulse shape, m the frequency-modulation index, f_m the bandwidth of each of the n_c channels being transmitted.

This necessitates the separation of the channels before frequency discrimination takes place.

The relation between signal-to-noise ratios in message and line is assumed to be the same as for standard frequency modulation;

$$\frac{S_l}{N_l\sqrt{2}}=\frac{1}{m\sqrt{3}}\frac{S_m}{N_m\sqrt{2}}.$$

The extra bandwidth B_0 is determined by the fact that the interference caused by one pulse on the following should be less than the effect of noise. Assuming that 60 decibels are required for the message signal-to-noise ratio, this means that

$$20\log e^{\pi B_0\tau}\geqslant 60,$$

τ being the pulse duration.

This gives

$$B_0\tau\geqslant 2.2.$$

TABLE 4

TRANSMISSION EFFICIENCY FOR PULSED-FREQUENCY-MODULATION SYSTEMS

m	Transmission Efficiency η			
	$n_c=1$	$n_c=20$	$n_c=100$	$n_c=400$
1	0.08	0.12	0.12	0.12
5	0.07	0.15	0.16	0.16
10	0.06	0.17	0.19	0.19
50	0.023	0.20	0.29	0.31
100	0.020	0.21	0.36	0.42

Taking the lower limit, the expression for the transmission efficiency is found to be

$$\eta=\frac{1}{8.8+2\dfrac{(m+1)}{n_c}}\,\frac{3}{2.76-\log m}.$$

The results are tabulated in Table 4 for various numbers of channels. In all cases, the frequency-modulation index should, of course, remain under a value leading to a signal-to-noise ratio so low that the relation between signal-to-noise ratios in message and line is no longer valid.

7. Conclusions

The foregoing examples give sufficient illustration of the usefulness of the transmission-efficiency coefficient as defined, together with its limitations. It will be seen that frequency modulation and pulse-time modulation behave in a very similar way; the transmission efficiency decreases as the bandwidth used for the same overall modulating bandwidth is increased. This is obvious when it is recalled that the transmission efficiency is nothing more, in this case, than the product of the ratio of bandwidths in message and line by the ratio of signal-to-noise ratios in message and line, the second factor being expressed in decibels. The increase in bandwidth is not sufficiently compensated for by the decrease in decibels of the signal-to-noise ratio in the line. This should not be interpreted to mean that an increase in bandwidth should not be considered for such transmission systems, as the gains in signal-to-noise ratio in the line may well be the decisive factor, as, for instance, in the case of radio links where a reduction in transmitter output is generally a very desirable feature.

Pulse-count modulation is shown to have a higher transmission efficiency than the previously mentioned systems. It should be noted, however, that the apparent constancy of this factor for various numbers of digits in the code tends to mask the fact that the amount of information in the received message is not transmitted with equal accuracy for these different codes. The distortion introduced by the quantization process depends on the number of digits used. Should it be required that the distortion be of the same order as the noise tolerated in the message (that is, 60 decibels below signal level), it can be shown that the number of digits should be approximately 11. Fewer digits will give higher distortion, but this can be tolerated in many cases, although the noise should remain at a much lower level to avoid its being bothersome during intervals between speech. A higher number of digits would be unnecessary, unless a higher signal-to-noise ratio is to be achieved in the message.

The last two modulation methods have been included to show that an increase in bandwidth may lead to an increase in transmission efficiency, contrary to what happens for frequency modulation and pulse-time modulation. This is the case for pulse-amplitude modulation with the assumptions made; other considerations, such as fading phenomena on radio links, will often rule out this kind of modulation. Pulsed-frequency modulation appears to be quite favorable when the transmission of a great number of channels is considered, provided the circuitry involved does not prove to be uneconomical. On the other hand, this same system would not be suitable for transmitting a message occupying a large bandwidth in itself.

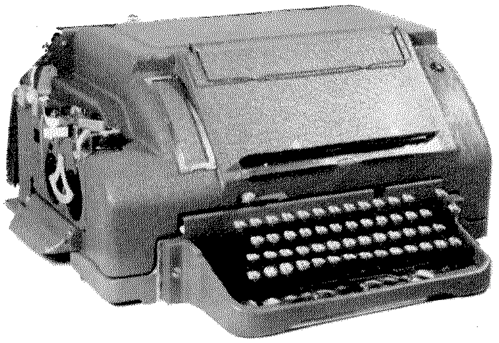
It should not be overlooked that transmission efficiency as defined is just one factor among many involved in the choice of a communication system to meet specific requirements. Prevalence of bandwidth considerations over signal-to-noise ratios, or *vice versa*, simplicity and reliability of equipment, both at transmitter and receiver, number of repeaters, degree of privacy required, cost of manufacture and maintenance; all of these are significant and cannot be included in an oversimplified factor of merit such as transmission efficiency.

There is, perhaps, no better conclusion than to quote Hartley, who stated,[1]

> It will, of course, be found that in very many cases it is not economically practical to make use of the full physical possibilities of a system. A criterion is, however, often useful for estimating the possible increase in performance which may be expected to result from improvement in apparatus or circuits, and also for detecting fallacies in the theory of operation of a proposed system.

Recent Telecommunication Developments

CREED No. 47 TAPE TELEPRINTER—Plans have been initiated for converting the British Inland Telegraph service to fully automatic switching. The introduction of nationwide automatic switching is also contemplated by the telegraph administrations of France, Holland, Belgium, and Denmark.



For such services to be economically successful, the teleprinters must not require operators in constant attendance. No time must be lost in receiving an answer to an incoming call or in confirming that a message has been correctly received. It must also be possible to transmit to any teleprinter with complete assurance that connection with the correct office has been made, that its printer is in a condition to receive traffic, and that there is an ample supply of paper tape that is feeding correctly. The teleprinters must also be capable of long periods of heavy service with minimum attention. To meet these needs, Creed and Company has introduced a new tape-model teleprinter, which embodies those units of the No. 7 page model that have proved entirely dependable under the most arduous conditions. Many novel features, which make this new machine particularly suitable for use in automatic switching networks, have also been included.

An "answer-back" unit provides confirmation that the correct office has been contacted.

The typewriter ribbon and roll of paper tape can be renewed without removing the printer cover or stopping transmission. The location of the printing point and of the tape tear-off position ensures good visibility and ease in handling incoming traffic. Should the paper tape break, become expended, or fail to feed forward, a tape alarm gives positive warning immediately and transmission is interrupted.

An orientation device, in association with high-grade signals from the striker transmitter fitted to the keyboard, furnishes a ready means of testing the operating margin of the instrument.

Maintenance is facilitated by unit construction and a lubrication system permitting over 300 hours continuous operation without attention.

A four-row keyboard can be supplied and incorporates a shift-lock mechanism as a safeguard against failure to operate the "figure" and "letter" shift keys. A three-row keyboard with double character keys or a commercial typewriter keyboard with automatic "case" change mechanism can also be provided.

A message tray may be used. It incorporates a station indicator and provides convenient means for the separation and storage of incoming and outgoing traffic.

The dimensions of the printer makes it well suited for use at manual switching positions.

• • •

ETCHINGS OF ERLANG—The Bureau de l'Union Internationale des Telecommunications published in 1935 the first of a series of etchings of noted contributors to the telecommunication sciences and arts. There has recently been added to this series a portrait of Agner Krarup Erlang.

Erlang is widely recognized for his mathematical work in the field of telephony, which is perhaps best epitomized by his formula for the distribution of telephone waiting times. Although most of his professional life was spent during the opening decades of the twentieth century with the Copenhagen Telephone Company, his interests were not restricted to telephony but embraced physics and electricity and among the arts, philosophy, history, and poetry.

All etchings in the series are still available from the Bureau at Effingerstrasse 1, Berne, Switzerland at 3 Swiss francs each, including mailing and packing. The etchings including margins are 23 by 17 centimeters (9 by $6\frac{5}{8}$ inches) on luxury paper. Portraits of the following may be ordered: Ampère, Baudot, Bell, Erlang, Ferrié, Gauss and Weber, Hertz, Hughes, Marconi, Maxwell, Morse, Popov, and Siemens.

In Memoriam

JOHN EDWARD KINGSBURY

John Edward Kingsbury died on November 4, 1948. At the time of his death in his 94th year, he was the oldest man in the world who had been concerned with the development of the telephone. He had been associated with many telephone pioneers including Alexander Graham Bell.

In 1883, Mr. Kingsbury opened a small office in London as the representative of the Western Electric Company. From this modest beginning was developed Standard Telephones and Cables, Limited, an associate company of the International Telephone and Telegraph Corporation.

Telephone equipment and exchanges were supplied and installed for the National Telephone Company and other companies. In 1897, he was instrumental in purchasing the North Woolwich works of Fowler Waring Cable Company, which is still a manufacturing unit of the Standard organization. Mr. Kingsbury retired from active management in 1906 but continued until 1925 as a director of Western Electric Company Limited. He was also actively associated with Damard Lacquer Company and with Bakelite Limited, with which the former company was merged.

Mr. Kingsbury joined the Society of Telegraph Engineers in 1887 and played an active part in developing that organization into the Institution of Electrical Engineers. He served as vice president, honorary treasurer, and as a member of the council of the Institution between 1900 and 1920.

His exhaustive treatise, "The Telephone and Telephone Exchanges—Their Invention and Development" was first published in 1915 and is still consulted as a standard work.

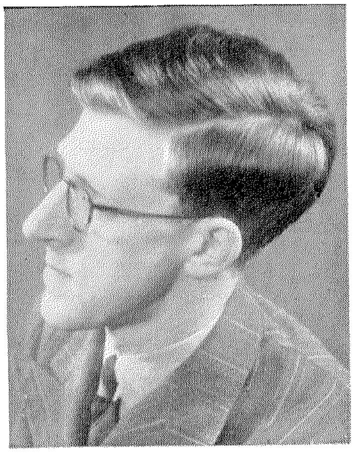
Contributors to This Issue

Richard F. Cleaver

Richard F. Cleaver was born near Liverpool, England, on July 12, 1912. He received the B.Sc. (Eng.) and M.Sc. (Eng.) degrees from Owens College in the University of Manchester in 1933 and 1934.

He joined the radio division of Standard Telephones and Cables Limited in London in 1934 and has specialized in radio direction finding. Until the outbreak of the recent war, he worked on medium and high-frequency manual direction finders for use at aeronautical ground stations, spending considerable time in South Africa on installation and experimental work in the field. Since 1939, he has worked mainly on the development of automatic direction-finding systems.

Mr. Cleaver is an Associate Member of the Institution of Electrical Engineers.



David L. Hollway

• • •

David L. Hollway was born at Ballarat, Victoria, Australia, in December, 1915. At Melbourne University, he received the B.E.E. degree in 1937 and the M.Eng.Sc. degree in 1939.

From 1940 to 1946, he served in the valve division of Standard Telephones and Cables Pty., Limited, handling engineering problems in the production of receiving, transmitting, and radar vacuum tubes.

In July, 1946, he joined the staff of the Electrotechnology Division of the National Standards Laboratory of the Council for Scientific and Industrial Research, where he is at present engaged in work on ultra-high-frequency measurements.

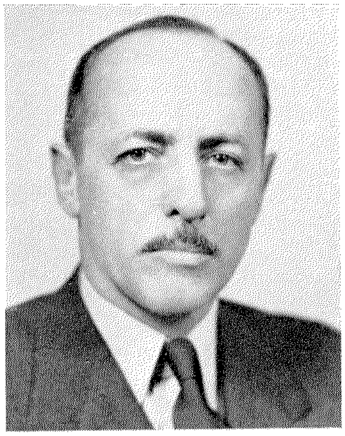
Mr. Hollway is an Associate Member of the Institution of Engineers, Australia, and of the Institution of Radio Engineers, Australia.

• • •

George Lewis was born in Media, Pennsylvania, on September 15, 1887. He attended St. John's College and Columbia University.

From 1910 to 1912, he served as a radio engineer in the U. S. Army Signal Corps. In 1912, he became an expert radio aide in the U. S. Navy, rising to the rank of Commander. In 1922 he joined the Crosley Radio Corporation as assistant to the president. He served as vice president and general manager of the Ken-Rad Corporation from 1925 to 1927 and filled similar positions in the Arcturus Radio Corporation from 1927 to 1935.

He joined the engineering staff of International Telephone and Telegraph Corporation in 1935 and is now an assistant vice president of that organization and of the Federal Telephone and Radio Corporation.



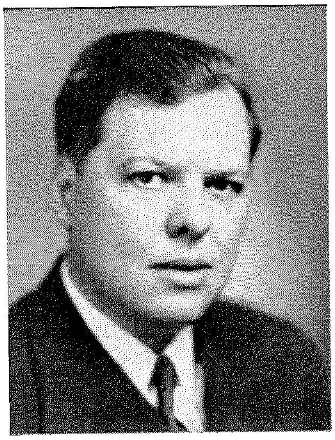
George Lewis

Mr. Lewis is a member of the Institute of Radio Engineers, Engineers' Club, Radio Club of America, and the Annapolis Yacht Club.

• • •

F. J. Mann was born in Hoboken, New Jersey, on April 27, 1906. He received the B.S. degree from Middlebury College in 1933. Postgraduate studies were in communications and journalism, the latter at Columbia University School of Journalism.

From 1925 to 1927, he was on the staff of *Radio Broadcast* magazine. In 1928, he was appointed radio editor of



F. J. Mann

Alexander W. Montgomery

Science and Invention. He served as a marine radio operator before completing college and after his post-graduate work he became a radio operator at broadcast station WCAX in Burlington, Vermont. From 1934 to 1940, he was engaged in newspaper work and free-lance writing. Mr. Mann then became assistant to the director of technical publications of Bendix Radio.

Joining the *Electrical Communication* staff as a writer in 1942, he is now manager of the Technical Publications Division of I.T.&T., managing editor of *Electrical Communication,* and editor of the handbook, *Reference Data for Radio Engineers.*

• • •

Alexander William Montgomery received the B.Sc. Tech. degree from Manchester College of Technology after having served for four years with the army during World War I.

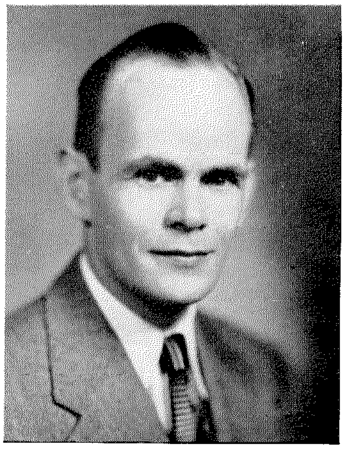
In 1921, he joined the Western Electric Company in London and became closely associated with the early development in Great Britain of repeaters and carrier systems. He was active also in the field of voice-frequency telegraphy. He is now technical director of Standard Telephones and Cables, Limited, and managing director and director of research of Standard Telecommunication Laboratories, Limited. During World War II, he was actively engaged on the development of the British defense teleprinter network and of a wide range of communication equipment. He served on a number of government committees and, in 1944, visited the U.S.A. as a member of a Ministry of Supply mission. For his war work, he was made an Officer of the Order of the British Empire.

For many years, Mr. Montgomery has been prominently identified with both the Comité Consultatif International Téléphonique and Comité Consultatif International Télégraphique, and represents the British telecommunications industry on a number of government and other committees.

He is a Member of the Institution of Electrical Engineers and of the American Institute of Electrical Engineers, and a Senior Member of the Institute of Radio Engineers.

• • •

Norman H. Young was born in Philadelphia, Pennsylvania, in 1913. He received the B.S. degree in electrical engineering from Pennsylvania State College in 1934 and the M.S. degree in 1935.



Norman H. Young

From 1935 to 1942, he was engaged in television engineering for the Philco Corporation and had charge of the transmitter of television station WPTZ.

In 1942 he became a department head in Federal Telecommunication Laboratories and during the war was largely concerned with the application of pulse-time modulation to military communication equipment. At the termination of the war, he was responsible for the engineering of the color-television transmitter for the Columbia Broadcasting System. He has done additional work on receivers and studio equipment for color television.

Mr. Young is a member of Eta Kappa Nu and a Senior Member of the Institute of Radio Engineers.

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For a biography and photograph of A. G. Clavier, see v. 25, p. 206; June, 1948.

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Compañía Internacional de Radio, S.A., Santiago, Chile

Radio Corporation of Cuba, Havana, Cuba

Radio Corporation of Porto Rico, Santurce, Puerto Rico[1]

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Mackay Radio and Telegraph Company, New York, New York[3]

All America Cables and Radio, Inc., New York, New York[4]

Sociedad Anónima Radio Argentina, Buenos Aires, Argentina[5]

[2] Cable service. [3] International and Marine Radiotelegraph services.
[4] Cable and Radiotelegraph services. [5] Radiotelegraph service.

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